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DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH

RADIO RESEARCH

SPECIAL REPORT No. 8

AN INVESTIGATION OF THE INTERFERENCE CAUSED BY TRANSMISSIONS FROM RADIO STATIONS

PART I. THE MEASUREMENT OF THE EQUIVALENT DECREMENT OF VARIOUS TYPES OF RADIO TRANSMISSIONS

BY

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PART II. THE STUDY OF THE INTERFERING PROPERTIES OF RADIO TRANSMISSIONS

BY

F. M. COLEBROOK, B.Sc., D.I.C., A.C.G.I.

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INTRODUCTORY NOTE

THIS Report describes an investigation commenced in April, 1923, under the direction of the Radio Research Board in response to a request from the Sub-Committee on Technical Wireless Regulations of the Imperial Communications Committee. This Sub-Committee, being engaged in drafting national wireless regulations which might form the basis of proposals for the 1927 meeting of the International Radio Telegraph Convention, asked the Board to consider the utility of certain figures for "equivalent decrement" as a basis for the classification of transmitting stations in respect of their interfering properties; "equivalent decrement" being a quantity defined in Section LXXXII of the Regulations of the Washington Draft Convention of 1920, and described more fully in Part I of the Report.

From the results of experimental measurements carried out by the National Physical Laboratory during the years 1923–24 it was concluded that the quantity "equivalent decrement" was not an adequate criterion of interference in radio communication. Some experiments were also made by Mr. F. E. Nancarrow, of the Post Office Engineering Department, which confirmed this conclusion. At this stage the Sub-Committee on Technical Wireless Regulations set up a special Sub-Committee on Interference with instructions to report on the subject of equivalent decrement and, if necessary or desirable, to present alternative proposals. On the basis of further experimental work by the National Physical Laboratory during 1925 this Sub-Committee confirmed the inadequacy of "equivalent decrement" and put forward alternative proposals based on the energy spectrum of the transmission. As it was apparent that considerable research on this subject would be necessary before any practically realisable specification of interfering effect could be formulated, the above Sub-Committee was

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absorbed into a special Committee set up by the Radio Research Board in 1926 to conduct this research.

During the years 1926–28 a considerable amount of theoretical and experimental work was carried out at the National Physical Laboratory, but this had failed to produce a practical and easily applied method of measuring interference. In the meantime, owing to various improvements in the technique of radio communication, the necessity for immediate legislation on the subject had decreased. Furthermore, largely as the result of the work carried out in this country and described in the present Report, the International Radio Telegraph Convention held at Washington in 1927 abandoned the use of the quantity " equivalent decrement " and made no alternative suggestion for defining interference. It was decided, therefore, to discontinue this rather unpromising line of research. In view, however, of the probable importance of the problem of interference in connexion with the future regulation of radio communications, it has been considered desirable to draw up the accompanying Report as a record of the progress made and a statement of the difficulties to be overcome.

E. H. SHAUGHNESSY,
Chairman of the Interference Committee.

DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH,

16 Old Queen Street,
Westminster, S.W.1.
November, 1928.

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AN INVESTIGATION OF THE INTERFERENCE CAUSED BY TRANSMISSIONS FROM RADIO STATIONS

PART I. THE MEASUREMENT OF THE EQUIVALENT DECREMENT OF VARIOUS TYPES OF RADIO TRANSMISSIONS

1. Proposed Classification of Radio Transmitting Stations and the Definition of "Equivalent Decrement"

In Section LXXXII of the Regulations of the Washington Draft Convention of 1920 it was proposed to classify radio transmitting stations in accordance with the type of waves they emit rather than in accordance with the type of apparatus used for producing the waves. The proposed classification of waves was as follows :—

Type A.—Continuous Waves.—Waves arising from oscillations which, after reaching the steady state, are periodic, i.e., the successive oscillations are identical.

Type A1.—Continuous Waves, key modulated.—Continuous waves of which the amplitude or frequency is varied by the operation of keying in telegraphic transmission.

Type A2.—Continuous waves modulated by audible frequency.—Continuous waves in which the amplitude or frequency is varied in a periodic manner at an audible frequency.

Type A3.—Continuous waves modulated by speech.—Continuous waves in which the amplitude or frequency is varied according to the characteristic vibrations of speech.

Type B.—Damped waves.—Waves consisting of successive wave trains in which the amplitude of the oscillations, after reaching its maximum, declines gradually.

These definitions do not refer to types of transmitting apparatus. They do not exclude, for example, from type A1 emissions from spark apparatus which produce real continuous waves. Within each of the above types it was proposed to classify the waves according to the degree of interference they produce in communication work, making use for this purpose of a quantity termed "equivalent decrement," which is derived in the following manner.

The resonance curve of the root-mean-square value of the current (I) in a receiving circuit is determined with respect to the

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TABLE 1.—*Proposed limiting values of equivalent decrement to be allocated to the various types of transmissions on different wave lengths*

Wave.		Type and Decrement.			
Frequency.	Length.	A1.	A2.	A3.	B.
∞ –300 kc./s. ..	0–1,000 m. ..	0·000 to 0·005	0·005 to 0·02	0·02 to 0·08	0·08 to 0·16
300 –105·3 kc./s. ..	1,000–2,850 m. ..	0·000 to 0·005	0·005 to 0·02	0·02 to 0·08	0·08 to 0·16
105·3– 37·5 kc./s. ..	2,850–8,000 m. ..	0·000 to 0·005 (hand) .. 0·005 to 0·02 (auto) ..	0·02 to 0·08	0·08 to 0·16	0·08 to 0·16
37·5– 0 kc./s. ..	8,000– ∞ m. ..	0·005 to 0·02	—	—	—

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natural frequency (f) of the circuit. The quantity δ defined by the equation :—

$$\delta = \pi \frac{f_1 - f_2}{f_r} \sqrt{\frac{I^2}{I_r^2 - I^2}}$$

is then determined from the resonance curve ; f_1 and f_2 being the two frequencies on either side of f_r (the frequency of the transmission) at which I is measured, it being conditionally accepted that the frequencies be selected over the range between the limits $0{\cdot}9 f_r$ and $1{\cdot}1 f_r$ and that I^2 shall not exceed $\frac{1}{2} I_r^2$. If the quantity so determined is found to vary with f, the largest value obtained should be taken. A repetition of the above measurement with a continuous wave source of sine wave radiation of the same frequency will give the value of δ for the receiving circuit. The difference between the two values of δ is then described as the " equivalent decrement " of the given transmission. On the basis of this definition the preceding table (Table 1) of permissible decrements for the various types of transmission mentioned above was formulated provisionally, and it was with the object of ascertaining the practical utility of the figures in this table that the first experiments were undertaken in 1924.

2. Methods of Measuring Equivalent Decrement

In a paper read before the Institution of Electrical Engineers in 1922, Erskine-Murray and Williams[1]* described an experimental method of measuring the decrement of the waves transmitted from a distant spark station. Using a frame coil receiver, the combined decrement of the transmission and the receiver was measured by observing the change in the capacity of the condenser required to detune the receiver by an amount sufficient to reduce the received current to $\frac{1}{\sqrt{2}}$ of its maximum value in the condition of circuit resonance. The received currents were not measured directly, but the ratio $\frac{1}{\sqrt{2}}$ was obtained by comparison with the current received on the tuned coil first, when it was set in the maximum position and secondly, when rotated through 45° from this position. The current was indicated either by a crystal detector and galvanometer coupled to the last audio-frequency stage of the receiver, or by the use of a thermionic voltmeter. In the discussion on this paper it was pointed out by Round that such instruments do not necessarily read R.M.S. values of current or voltage for damped oscillations, and that since in the two sets of measurements the receiving loop is respectively tuned and distuned, it must not be assumed that the envelope of the wave-form of the received current

* These numbers refer to the bibliography on page 40.

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is the same in the two cases. It is thus doubtful if equality of rectified signal current in the two observations implies equality of the R.M.S. values of these currents, and the corresponding error would vary with the wave-damping and type of transmission and is, in any case, unknown. The decrement of the receiving circuit was measured by Erskine-Murray and Williams by separate measurement of its constants with the aid of a local source of continuous oscillations, all conditions being maintained as far as possible identical with those in the previous measurements. The only decrement value given in the paper refers to observations on the damped wave " spark " transmissions from Paris, the value obtained on a wave-length of 2,600 metres being 0·16. It is not stated if this referred to a steady dash or to the transmission of Morse signals.

For the determination of the equivalent decrement of a transmission as defined in the previous section, it is clear that the principal requirement is an instrument capable of measuring small radio-frequency currents in the receiving circuit, or of the small radio-frequency potentials of the order of from 0·001 to 0·1 volts which result across the tuning condensers of such a circuit. The instrument should not add substantially to the decrement of the circuit to which it is attached, its calibration must remain reasonably constant, and by no means least important, the indications of the instrument must follow a square law over its whole working range so that it will read the mean square value of a current of any wave form. With the advance in the technique of radio-frequency measurements it is evident that the valve voltmeter in one or other of its several forms meets the above requirements, and it was with an early pattern of such an instrument that measurements of equivalent decrement were begun at the National Physical Laboratory in 1923. From the commencement, it was considered desirable to plot complete resonance response curves in these measurements rather than to limit the observations to specified current values. To obtain the necessary scale range for this purpose, combined with the square-law condition, necessitated the use of a fairly sensitive reflecting galvanometer to indicate the change in anode current of the valve. By means of this type of instrument a reading proportional to the square of the voltage induced in a receiving circuit by the station being measured can be obtained. A curve showing variation of this reading with a suitable variation of the tuning condenser on either side of resonance can be plotted. A similar curve can be obtained for a local continuous wave source of the same wave-length or for continuous waves transmitted from the same station. Finally, the receiving circuit can be calibrated for wave-length over the range of tuning condenser employed in the first two curves, by means of a local oscillator or wave meter. These three curves will give all the information required for the determination of the equivalent decrement of the transmission.

The type of receiving circuit used varied with the conditions of measurement. Thus the stations at Leafield, Northolt and Ongar

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gave ample signal strength on a 100-foot aerial with parallel condenser tuning, but unless specially arranged transmissions were sent from these stations for the purpose of the measurement, the aerial could not be used owing to the impossibility of obtaining the C.W. and wave-length calibrations, since the stations worked almost continuously. The readings for these stations were, therefore, taken on a frame receiving coil which could be set in the position of zero reception for the purpose of calibration on the local oscillator. It was found necessary to use a radio-frequency amplifier in conjunction with the frame to make the received signals large enough for measurement.

Apart from this difficulty of the local oscillator calibration, however, the use of the aerial and the direct measurement of the signal e.m.fs. are to be preferred. A special advantage of the latter is that by inserting a suitable coil in the anode circuit of the measuring valve and coupling this to the aerial tuning inductance, the decrement of the whole receiving circuit can be reduced to as low a figure as 0·002 without any instability.

In the case of some later measurements made on the transmissions from the North Foreland " spark " station it was necessary to employ portable apparatus, and the limitations of such apparatus are referred to later.

3. Results of Measurement of Equivalent Decrement

(*a*) *Experiments made at the National Physical Laboratory*

With the aid of the apparatus described above a number of measurements were carried out at the National Physical Laboratory on various types of wireless transmissions. For the most part normal commercial transmissions were made use of for this purpose but a local transmitter at the Laboratory was also employed. A summary of the results obtained is given in the following Table 2.

Table 2.—*Summary of results of measurements of equivalent decrement made in* 1924 *at the National Physical Laboratory on various types of transmissions*

Station and Wave-length.	Type of Transmission.	Equivalent Decrement.	Permissible Equivalent Decrement proposed in Section LXXXII.
Northolt (GKB), 6,950 metres	Keyed C.W., hand speed; two wave-lengths	0·003	0·000–0·005
Ongar (GLB), 3,840 metres	High speed. Automatic C.W.	Less than 0·001	0·005–0·02

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TABLE 2—*continued*

Station and Wave-length.	Type of Transmission.	Equivalent Decrement.	Permissible Equivalent Decrement proposed in Section LXXXII.
Leafield (GBL), 8,570 metres	Keyed C.W., hand speed ; two wave-lengths (8,528 and 8,612 metres)	0·011 to 0·024, depending on the ratio of I^2 to I^2_r	0·005–0·02
N.P.L. (5HW), 762 metres	Spark—		
	Long dash ..	0·05	0·08 to 0·16
	Dots at hand speed	0·05	0·08 to 0·16
N.P.L. (5HW), 762 metres	Interrupted C.W.	0·006	0·005–0·02
London (2LO), 365 metres	C.W. Modulated by speech and music	Less than 0·001	0·02–0·08
Chelmsford (5XX), 1,600 metres	C.W. Modulated by speech and music	Less than 0·001	0·02–0·08

In the case of telegraphic working the use of the normal commercial transmission of the station is not very suitable for accurate measurement since the galvanometer deflections show rapid fluctuations due to variations of mean intensity caused by the Morse keying. The readings recorded were mean values. By measurements made on a special transmission from the Laboratory station (5HW) it was found that a regular train of dots at hand speed gave a constant deflection, and it will be noticed that the equivalent decrement measured in this manner agreed with the value obtained on a steady dash on the spark transmitter. The actual curves obtained in this latter case are shown in Fig. 1. In the case of the observations made on the arc transmissions from Leafield (GBL) measurements were obtained on Morse signalling and on both the continuous marking and the continuous spacing wave, as shown in Figs. 2 and 3. For a curve of the type shown in Fig. 2 the equivalent decrement measured as defined above is not constant but appears to increase with $I_1^2\ I_2^2$ up to $I_1^2 = \frac{1}{2} I_2^2$.

(b) Experiments made at the North Foreland Station

With a view to obtaining further information on the application of equivalent decrement to spark transmitters, some measurements were carried out on the transmissions from the Post Office "spark" station at North Foreland (GNF) in December, 1925.

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Some preliminary observations carried out on these transmissions showed that, owing to the comparatively low power of the station and the interference from other stations on the same

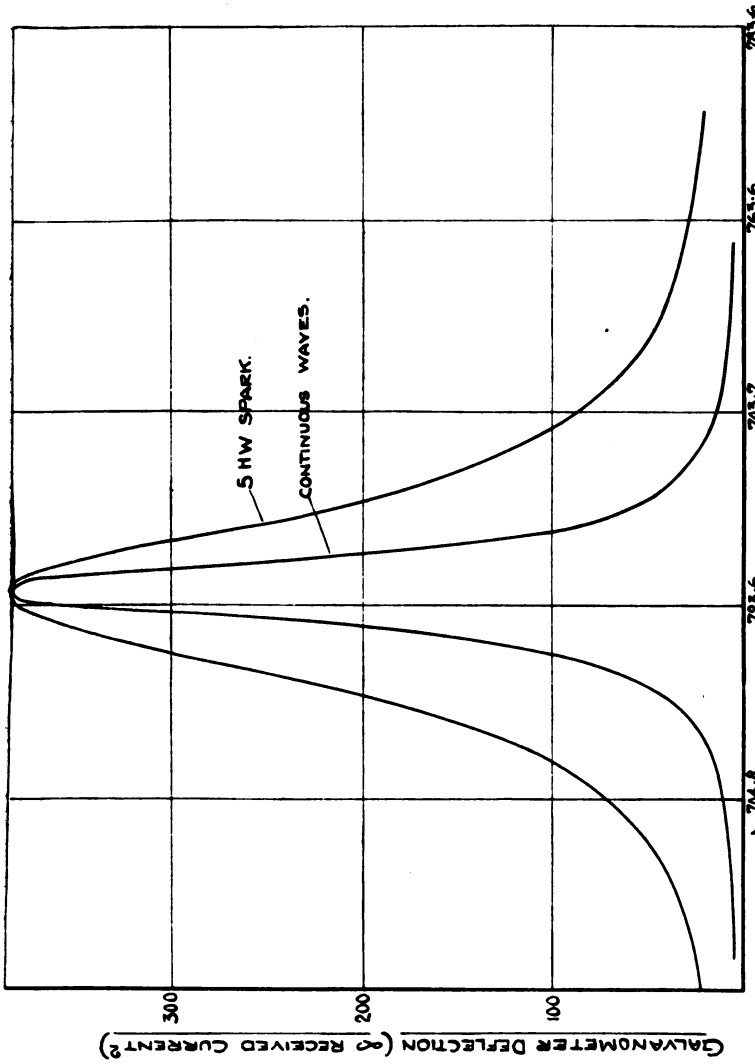

FIG. 1.—*Resonance curves showing current in a receiving circuit for a steady dash from a spark transmitter at the National Physical Laboratory and from a local source of continuous waves.*

wave-length, it was impracticable to make any accurate measurements at Teddington, about 90 miles distant. It was, therefore, decided to modify the apparatus previously employed so as to make it fairly portable, and then to use it within a comparatively short distance of the transmitter.

A frame coil, 2 feet square, was employed with suitable tuning condensers, and the potential difference across these was measured

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with a simple rectifying valve arrangement with the steady anode current balanced out. On account of portability it was necessary to use a pointer galvanometer; and it was found that the most sensitive of these suitable for the purpose did not permit of the

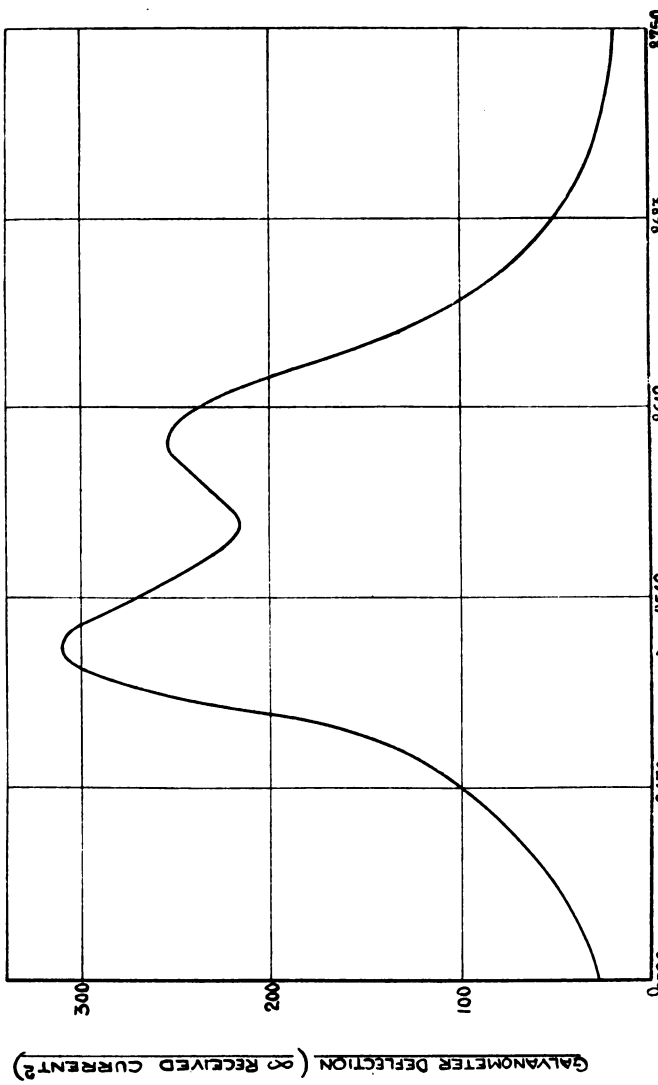

FIG. 2.—*Resonance curve of received current from Leafield (arc) transmitter obtained during the sending of Morse.*

valve being worked within the range over which the change in anode current of the rectifying valve was proportional to the square of the voltage applied to the grid. This meant that the instrument would not read true R.M.S. values for trains of damped waves. It was found that the response of the instrument was such as to

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make the resonance curve steeper than it should be, corresponding to a lower value of equivalent decrement, but in view of the urgency of forming some definite opinion as to the utility of the quantity "equivalent decrement," it was considered expedient to make the approximate measurements possible with the apparatus.

The measurements were made in the lighthouse at North Foreland, the distance from the frame to the nearest point of the aerial being about 100 yards. At this distance the resonant potential

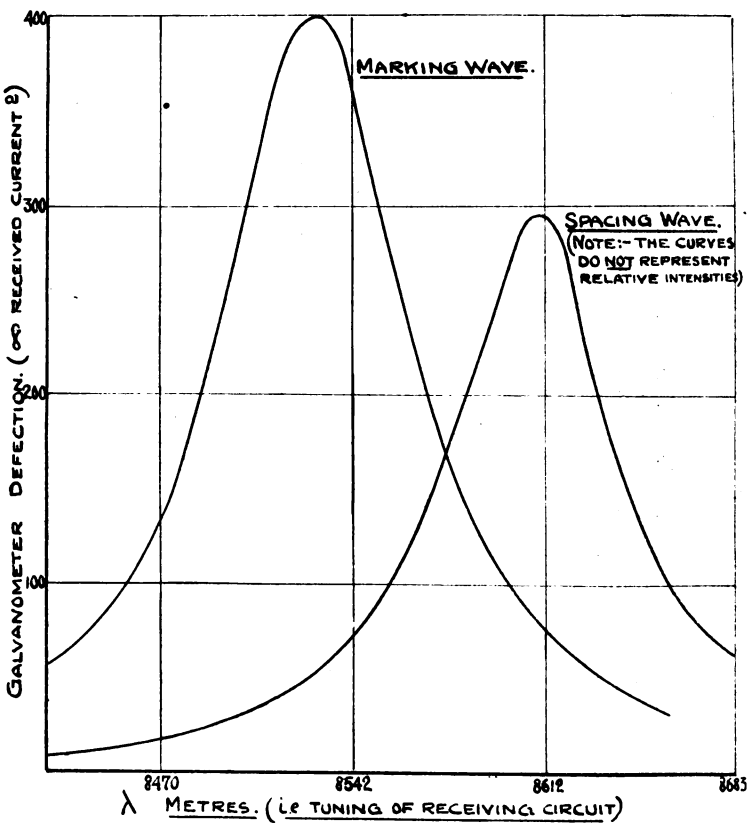

FIG. 3.—*Resonance curves of received current from Leafield for a steady dash on both marking and spacing waves.*

difference across the frame terminals was over 4 volts. As full scale deflection on the galvanometer was obtained for about 1·2 volts, this made it possible to measure the peak of the resonance curve with the frame turned out of the maximum position, and so increase the relative sensitivity of the instrument for the lower portions of the curve.

It is realised that at such a short distance from the transmitter the effective wave-form of the radiation may be somewhat modified

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by the presence of the induction field of the aerial. This again is a somewhat obscure point that must be left for future investigation ; but it is anticipated that the error introduced by neglecting the induction field will not be serious. A careful test carried out with the aerial disconnected showed that there was no detectable field from the primary or intermediate circuits of the transmitter.

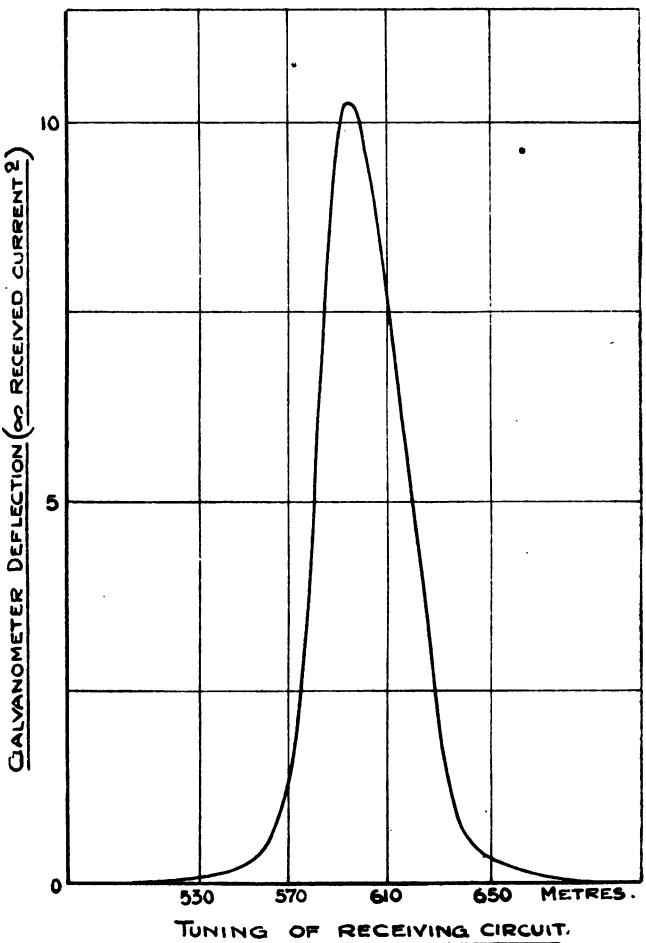

FIG. 4.—*Resonance curve of received current from North Foreland (spark) transmitter taken on a local loop receiver.*

The resonance curve of the receiving circuit when using a continuous wave source showed that the decrement of the receiving coil was fairly low, its actual value being 0·018. On the North Foreland station, the resonance curve obtained with the intermediate circuit in position, as normally used at that time, is shown in Fig. 4. It is at once evident that the transmission shows a

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serious equivalent decrement, its calculated value being 0·24. On either side of this main peak on 600 metres, the curve falls to the zero line but, by careful search, a second peak was discovered on a wave-length of about 760 metres, the height expressed as the square of the voltage of this small peak being about 0·0034 of that of the main peak. Although repeated search was made, no signal voltage could be detected in either this or any of the other tests, on wave-lengths corresponding to the second and third harmonics.

With a view to obtaining some further information about this second peak, the couplings between the primary and secondary, and between the secondary and aerial circuits of the transmitter were reduced slightly, such retuning as was necessary being made to bring the aerial current back to a maximum. By repeating the measurement of the resonance curve, it was found that the small peak had increased considerably in amplitude and that it now occurred at a shorter wave-length. After a second reduction of both couplings as above, the curve shown in Fig. 5 was obtained, with the peaks occurring at still closer wave-lengths. It would appear from these experiments that the existence of the second peak was due to the coupling of the circuits being too close, and that if this could be sufficiently reduced and a means of tuning each of the three circuits could be found, the two peaks could be merged into one. It is to be observed that the reduction of the couplings resulted in a slight increase in aerial current from 6·0 to 6·5 ampères under stable conditions of working. The calculated equivalent decrement from the curve in Fig. 5 vary from 0·09 to 0·17 depending upon the portion of the curve chosen for the calculation, the highest values being obtained at a part of the curve which embraces the second peak. It is evident that this value takes no account of the height of this second peak, and thus is not, therefore, a measure of the amount of interference caused by it.

Similar measurements were carried out with the intermediate circuit at the transmitter removed, and it was found that, while two peaks were obtained when the coupling was close, these could easily be reduced to one without serious reduction of the aerial current. In this latter condition the equivalent decrement of the transmission was found to have a mean value of 0·17. Since the merging of the two peaks into one by reduction of the coupling did not materially increase the equivalent decrement of the main peak, it is to be expected that if, with the intermediate circuit in place, the couplings could be reduced below the values used for Fig. 5, a value of the equivalent decrement of the order of 0·1 could be obtained. It would, furthermore, appear that it is only in this condition that the use of the intermediate circuit would be of any material advantage in reducing the interfering properties of the transmitter.

As was pointed out above, the inherent error of the valve voltmeter as used in these tests would have the effect of reducing the

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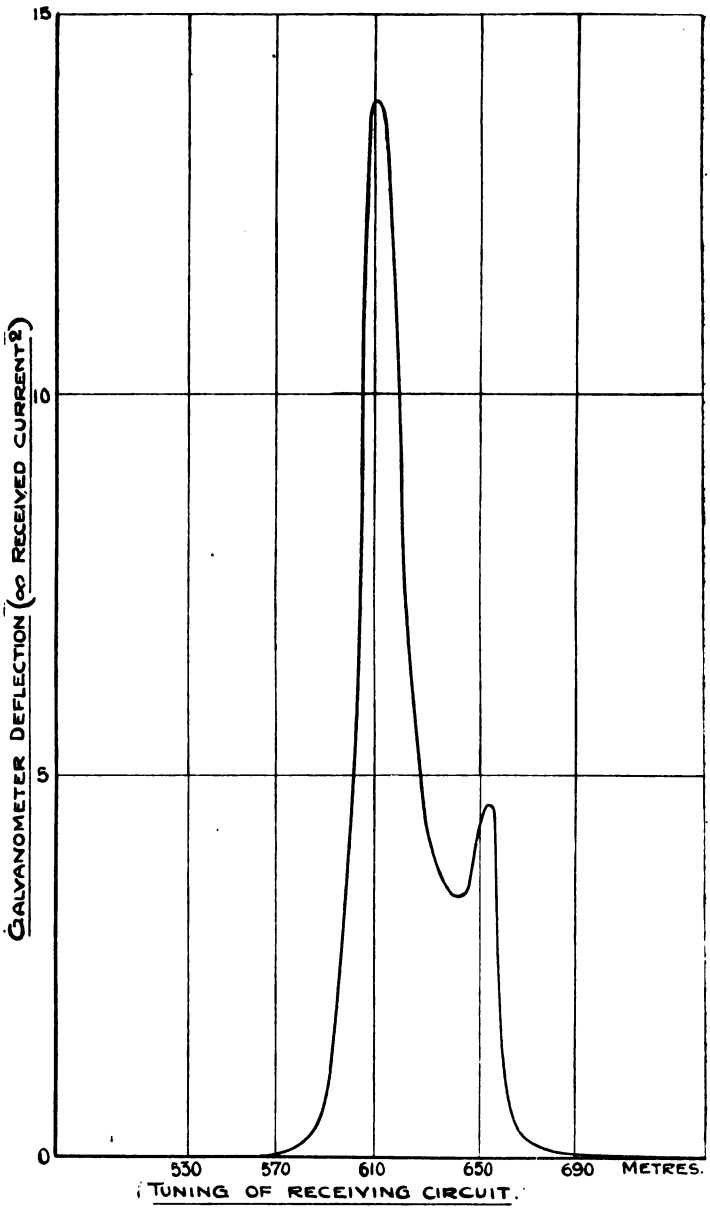

FIG. 5.—*Measurements in Figure 4 repeated with closer coupling of the circuits in the transmitter.*

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equivalent decrement values obtained. These are therefore to be considered as minimum values for the transmitter. In view of this fact, it is interesting to recall that the equivalent decrement given by the spark transmitter, as installed and measured at Teddington in 1924, was 0·05 (see section 3 (*a*) above). It is also to be recalled that, as stated in Table 1, the proposed permissible range of equivalent decrement for this type of transmission was 0·08 to 0·16.

4. Discussion of above Measurements and Conclusions to be drawn therefrom

From the experimental results given in the last section it will be seen that, in the case of the transmissions from arc and spark stations, the measured values of equivalent decrement are in fairly good agreement with the values originally proposed and detailed in Table 1. In the case of the spark transmissions, while the allocation of an equivalent decrement value may thus prove to be a minimum condition to be satisfied, it is considered to be inadequate for the following reasons, which were brought to light by the experiments carried out at North Foreland.

(*a*) The definition of equivalent decrement takes no account of possible additional peaks in a resonance curve on wavelengths differing by more than 10 per cent. from that of the principal peak ; although any such peaks, whether due to harmonics or to coupling conditions, might cause serious interference.

(*b*) In the event of the resonance curve showing two or more peaks in close proximity, then the highest decrement value, obtained from the lower part of the curve covered by both peaks, will not necessarily take account of the height of the subsidiary peak, although this would obviously affect the interference caused by the transmission.

In the case of continuous wave transmitting stations, the results quoted in the last section showed that there is no appreciable equivalent decrement arising from the modulation by telephony or Morse keying, except where Morse signalling is effected by a key controlled change in frequency. This, however, is quite inconsistent with actual experience of interference, for it is found that high speed telegraphy by key controlled amplitude modulation of single frequency waves will give rise to " key-click " interference in circuits tuned to neighbouring frequencies, which interference is generally considered to be more serious than that due to frequency change keying. The appreciable equivalent decrement associated with the latter system is, therefore, misleading as a means of rating its interfering effects.

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Further, it is a fact of experience that the degree of interference caused by medium wave-length telephony transmissions (e.g., 5XX), particularly in the case of telephone reception on frequencies differing by 10,000 cycles or less from that of the interfering station, is considerably greater than that which would be inferred from the very low equivalent decrement of such transmissions.

It is, therefore, to be concluded that equivalent decrement measurements will not serve as an equitable or effective check on C.W. transmissions, keyed, interrupted or modulated, for the following reasons :—

(*a*) Measurements have shown that the keying of single wave-length C.W. stations produces no appreciable effect on the resonance curve, but it is a fact of experience that it does give rise to interference on neighbouring wave-lengths.

(*b*) Stations of which the keying changes the wave-length would have an equivalent decrement larger than is actually representative of the interference caused by such transmissions.

(*c*) In the case of telephone transmissions in which the modulation percentage is small the measured equivalent decrement is practically zero, since the modulation only very slightly affects the root-mean-square value of the received current. This quantity cannot, therefore, represent the interfering properties of such transmissions.

(*d*) Finally, with all the above types of transmissions the equivalent decrement definition entirely ignores all the harmonics of the transmission, and these harmonics are known in some instances to be the most serious factor of the interference produced by the transmitter.

While it is to be concluded from the above considerations that as a general and adequate criterion of interference the quantity " equivalent decrement " is unsuitable, it has been suggested that it might possbly form a useful minimum condition for the basis of regulations on interference from spark stations, provided the restriction as to the $\frac{1}{2}I^2$ value were removed and provided that a further regulation restricted the emission outside the $\pm$ 10 per cent. band to some specified intensity.

5. Possible Alternative Telephonic Methods of Measuring Interference

The failure of the method described in Section 2 to discriminate between continuous oscillations on the one hand and either high-speed Morse transmission or telephony on the other, indicates that the interference experienced in practical communication is due to transient currents or currents of a kind that will not be recorded by

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a root-mean-square measuring instrument of the type contemplated in the equivalent decrement definition. This can be illustrated very simply in the case of a pure tone modulated carrier wave. Suppose the current be of the type

$$i = \hat{\imath}\ (1 + k \sin mt) \sin \omega t$$

$\omega/2\pi$ being a radio frequency, $m/2\pi$ an audible frequency, and k the modulation percentage. The square of this current is

$$i^2 = \frac{\hat{\imath}^2}{2}\left(1 + \frac{k^2}{2}\right) + \frac{\hat{\imath}^2}{2}\left(2k \sin mt - \frac{k^2}{2} \cos 2mt\right)$$
$$- \frac{\hat{\imath}^2}{2}\left(1 + \frac{k^2}{2} + 2k \sin mt - \frac{k^2}{2} \cos 2mt\right) \cos 2\omega t.$$

Of the terms on the right-hand side, the first group is a continuous current, and is *the only one which would be recorded by a root-mean-square measuring instrument.* In fact

$$\text{Mean value of } i^2 = \frac{\hat{\imath}^2}{2}\left(1 + \frac{k^2}{2}\right) = \mathrm{I}^2\left(1 + \frac{k^2}{2}\right).$$

The second group would not affect the measuring instrument but, being audible frequency currents, would obviously give rise to interference in telephone reception. The third group, consisting of radio frequency currents, would not affect the measuring instrument and would not cause interference in telephone reception.

Note further that if k is 0, the mean value of i^2 is I^2, which is less than before by the factor $\left(1 + \frac{k^2}{2}\right)$. Thus if k is 30 per cent. the mean value of i^2 is only increased by $4\frac{1}{2}$ per cent. by the modulation, which small variation is certainly no measure of the interference produced by the low frequency group of currents (which are proportional to k).

It is not suggested that this is an exact representation of an interference effect, but it serves to show that interference may and probably will arise from factors of which the equivalent decrement root-mean-square measurements do not take proper account.

In view of the above it was considered that some telephone audibility method might be more effective in estimating interference. In one of the methods investigated with this object in view, two screened variable mutual inductances were employed, either of which could be connected in turn between the aerial or frame coil and the receiver by the operation of a change-over switch. These inductances were identical in design and construction so that when they were both tuned and set at equal couplings, equal signals would be obtained in both positions of the change-over switch. By detuning the circuits connected to one of the mutual inductances the strength of signal received in the telephone is reduced, and this can be balanced against the signal received in the other positions of the switch with its corresponding circuits still

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tuned but with the couplings decreased by a suitable amount. Proceeding in this manner the decrease in signal strength produced by distuning one circuit could be measured by the diminution of coupling in the other circuit.

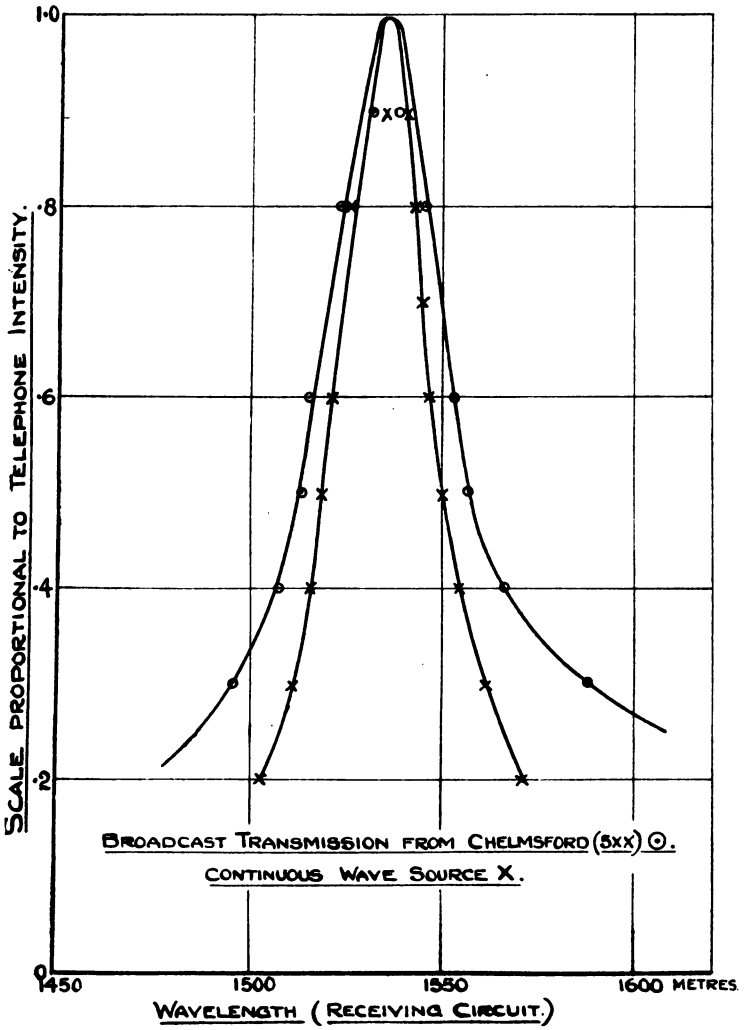

FIG. 6.—*Resonance curves illustrating the interfering effect of a transmission from Chelmsford (5XX) as measured by an audibility method.*

As an alternative to the use of variable mutual inductances, the two receiving circuits could be connected up through an audio-frequency amplifier giving a known voltage amplification variable over the range of 1 to 100.

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Both the above methods have been set up and tested in the experimental stage. In the case of high-speed Morse and of broadcasting transmission on a wave-length of 360 metres, no increase in the effective decrement could be observed over that obtainable on continuous waves. Some measurements made on the longer wave broadcast transmission from Chelmsford, however, did show a definite difference between the interfering effect of this station and that of a pure continuous wave, the results being illustrated by the typical curves given in Fig. 6. It was also observed that the detuning of the receiving circuit produced a very marked difference in the quality of the telephony which considerably increased the difficulty of establishing a telephone balance. This rather serious limitation combined with the somewhat low accuracy always associated with telephone audibility measurements, led to the suspension of these experiments while other methods of attacking the problem were being investigated. A description of these methods is embodied in Part II of this Report which deals with the further investigation of the whole problem of interference from wireless transmissions from a more general point of view.

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PART II. THE STUDY OF THE INTERFERING PROPERTIES OF RADIO TRANSMISSIONS

6. Proposals for a Criterion of the Interfering Capability of a Transmission

It has been shown in the preceding sections that "equivalent decrement" cannot be regarded as a satisfactory measure of the interfering capability of a wireless signal transmission. It now remains to outline proposals for a theoretically valid measure of interfering capability.

Any wireless transmission is the controlled radiation of energy through space, and the element of control which is essential for signalling purposes makes it impossible to restrict the energy radiated to a single frequency. In fact the only possible single frequency radiation is one of pure sine wave form of which the period and amplitude are constant, i.e., $\hat{e} \sin \omega t$ where t extends from $-\infty$ to $+\infty$. This, even if realisable, would be useless for signalling purposes. As distinct from this, any radiation which can be used for signalling purposes will of necessity have an imperfectly periodic character with a controlled variation of amplitude or frequency or both. It is a matter of common knowledge that such a disturbance is associated with a continuous distribution of energy with frequency (see Appendix 1). Theoretically the distribution will extend over the whole range of frequencies between zero and infinity cycles per second. In practice, however, the bulk of the energy carried by such a radiation will be associated with a bounded frequency zone in the neighbourhood of the principal frequency of the transmission.

The extent to which such a transmission will produce an undesired response in a circuit intended to receive some other transmission will obviously depend on the selectivity of the receiving circuit (i.e., its frequency-response characteristic) and on the energy-frequency characteristic of the transmission. It appears, therefore, that the energy-frequency or power-frequency characteristic of a transmission is the natural basis for any criterion of its interfering capability.

As an example of the energy distribution of the type of radiation involved in wireless signalling, consider a 1,000 KW. station which transmits a single Morse dot at a nominal frequency of 15,000 cycles per second at, say, a speed corresponding to 55 words per second (i.e., about 250 complete oscillations to a dot). For simplicity neglect the initial and final transient states of the transmitting system and assume that the radiated field at some given point can be represented by $e = \hat{e} \sin 30{,}000\pi t$, for all values of t from $t = 0$ to $t = 1/60$ second, and $e = 0$ for all other values of t. Fig. 7 shows the spread of the energy of a disturbance of this type over the spectrum in the neighbourhood of the mean frequency. (The

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figure actually corresponds to fifty complete oscillations which gives a more pronounced spread and is, therefore, better for illustration.) It can be shown that the ordinates corresponding to 20,000 and 20,200 metres, for the case of 250 oscillations, are in the ratio of 100 to 1·6. This means that if the radiated energy associated with a frequency band of width $\delta\omega$ at 20,000 metres is 1,000 kilowatt seconds, then 16 kilowatt seconds will be radiated at frequencies within a band of the same width at 20,200 metres.

In practice, of course, a transmission is not a single pulse as in the above example. In some cases (e.g., C.W. Morse transmission) it will consist of a sustained succession of such pulses. In others (e.g., radio-telephony) it will consist of a sustained train of carrier waves with an amplitude modulation of more or less complicated and continuously changing wave-form. Any isolated period of the transmission of finite duration can, however, be regarded as a pulse in the accepted mathematical interpretation of that word. It is inherently probable on physical grounds, and a theoretical investigation also supports the view, that if the duration of the isolated sample of the transmission be long enough, the ordinates of its energy-frequency characteristic will be proportional to the duration. If this condition be fulfilled, then the isolated element of the transmission can be regarded as a fair sample of the whole transmission from the interference point of view. The ordinates of the energy-frequency characteristic can then be divided by the time of duration of the sample, giving a mean power-frequency curve which can be regarded as characteristic for that particular transmission.

It is proposed that the mean power-frequency characteristic of a transmission, defined in the above manner, is the proper basis for the specification of the interfering capability of a transmission. This is in agreement with the view expressed in the pamphlet listed as No. 4 in the Bibliography, by Chester Snow (of the Bureau of Standards) :—

> " Undoubtedly a thoroughgoing spectrum analysis of the radiation when properly interpreted ought to furnish a logical basis for rating the interference producing qualities of a sending station."

On this basis there would be three separate elements involved in the qualification of any actual transmission of a given type in respect of its interfering capability.

(1) The determination of the ideal mean power-frequency characteristic, or, more shortly, the mean-power spectrum, of the given type of transmission.

(2) The determination, possibly by reference to an actual transmission considered to be good of its type, of the tolerance necessitated by material limitations and the existing state of the technique of the art.

(3) The determination by direct measurement of the actual mean power spectrum of the given transmission

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It is not too much to say that at the present time none of the evaluations involved in the above three steps can be carried out, except possibly the first. In view of that fact the present publication may be considered premature. In reply it may be urged that the matter is an important one and will increase in importance as

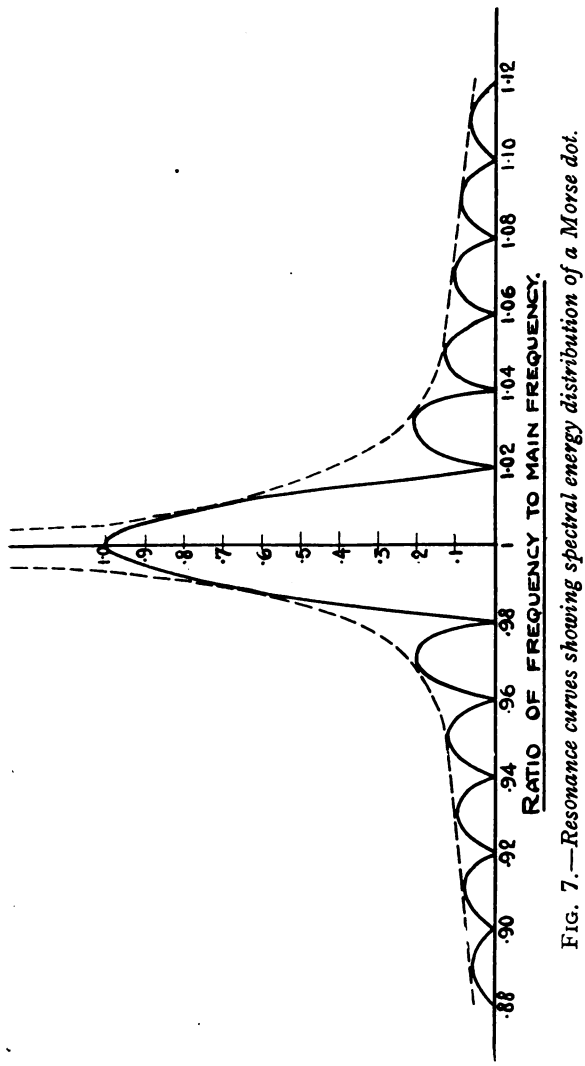

FIG. 7.—*Resonance curves showing spectral energy distribution of a Morse dot.*

time goes on and transmissions are multiplied. It is, therefore, very desirable that there should be a general understanding of the problems involved. The object of the present publication is to state these problems and to record what has so far been done toward their solution. The three elements detailed above will now be considered separately.

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7. Determination of the Ideal Mean-power Spectrum of a Transmission

The mathematical basis for this process is outlined in Appendix No. 1. It is derived for the most part from the paper No. 2 listed in the Bibliography (Appendix No. 4). Its application to any concrete case will obviously be exceedingly laborious. The simplest case will be that of pure C.W. Morse transmission. The calculation for a single Morse dot is given in detail, and the same process would be applicable to any succession of dots and dashes constituting an

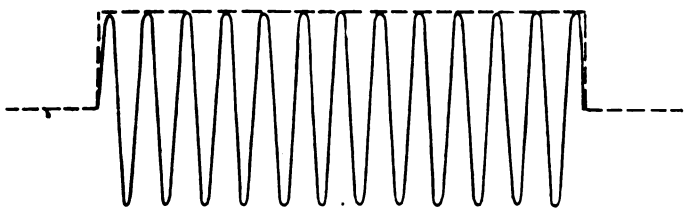

Fig. 8.—*Morse dot high-frequency signal, with square envelope.*

assumed sample of the transmission. It must be pointed out, however, that a practical difficulty arises at this point. The single dot considered as an example is assumed to have a square envelope as in Fig. 8. In practice the characteristics of the transmitting system will impose initial and final transient states which will result in an envelope of some shape such as that of Fig. 9. Now the

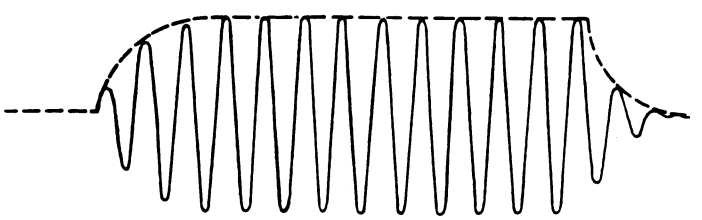

Fig. 9.—*Morse dot high-frequency signal, showing initial and final transient states.*

initial and final states play an appreciable part in determining the power spectrum of the disturbance, and very little is known at present as to the nature of these initial and final transients in actual transmitting systems. Until data on this matter are available complete calculations, even for C.W. Morse transmissions, cannot be carried out.

In the case of telephony, the difficulties are even more serious, owing to the complexity of the amplitude variation. In fact, it seems very improbable that it could ever be subject to computation

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of an exact character. However, if data were obtained as to the power spectrum of various types of modulation, e.g., speech, music, etc., it might be possible to form estimates of the limiting power spectrum of such telephony transmissions. A considerable volume of work is at present being carried out by commercial and national research laboratories on this subject.[12]

A great deal more could be said on the present aspect of the matter, but the above is sufficient to indicate the nature of the calculations involved and the considerable difficulties that have still to be overcome before they can be satisfactorily carried out.

8. Tolerance allowable to Actual Transmissions

This is the legal, commercial and international aspect of the matter and as such cannot here be fully discussed. Moreover, it obviously waits on a complete solution of the other two elements of the problem, which is very remote at present. Probably the most important single point for consideration would be whether permissible deviations from a standard should be expressed as absolute power or on a proportional basis. From the point of view of the interference actually experienced by a receiving system, the absolute magnitude of the interfering energy is the important matter, but an absolute specification would, of course, be much more drastic in effect on a high power than a low power station.

A further point involved is that not all forms of radiant energy are equally detrimental from the point of view of interference. In general, for instance, interference due to appreciable harmonics in the main frequency of the transmission will be more serious than that occasioned by the modulation of the main frequency.

In any case it must be recognised that no signal communication can be effected on a single frequency, even in the total absence of harmonics, and that any regulation restricting a station to a single frequency would, in effect, require the station to cease operations altogether.

9. Determination of the Mean-power Spectrum of an Actual Transmission

The mathematical basis of this part of the whole problem is outlined in Appendix No. 2. It will be seen that there are two main divisions of this part, namely (*a*) the determination by measurement of the mean-power spectrum of the response of a receiving system to a given transmission, and (*b*) the elimination from these experimental data of the known characteristics of the receiving circuit, leading to the mean-power spectrum of the transmission itself. These two parts will be taken separately.

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(*a*) *Determination of mean-power spectrum of response of a receiving system*

It will be seen from Appendix No. 2 that the measurement involved is the determination of the integral with respect to time of the electric energy produced in a tuned receiving circuit by the given transmission. The time of integration must be sufficiently long to satisfy the condition that the magnitude of the integral is proportional to time. To avoid the necessity for special controlled transmissions, the isolation of the sample of the transmission for measurement purposes can be effected by the closing and opening of the receiving circuit rather than by the starting and stopping of the transmission. This procedure will be justifiable as long as the time of integration is sufficiently long to render negligible the effect of the initial and final transient states in the receiving system, a condition which is included in that already stipulated. The measurement would require to be made at a sufficient number of receiving circuit tuned frequencies over the effective frequency range of the transmission to enable a curve to be drawn connecting integral mean-energy response with frequency.

The energy response in the receiving circuit will be a radio-frequency variation of more or less complicated wave-form. There is at present no known method of measuring the integral of such a quantity over appreciable periods of time. Section 10 contains a brief account of methods which have been tried at the National Physical Laboratory but without success. They are recorded simply for the information of any future experimenters on this subject.

A further point should be noted in connexion with the experimental side of this work. Apparatus of the type required will not be able to discriminate between the energy received from any given transmission and that derived from all other transmissions in simultaneous operation, nor can the contributions from all other existing transmissions be eliminated by a separate measurement in the absence of the transmission under investigation, since the energy derived from the sum of a number of separate currents is not the sum of the energies associated with each. This requires that the measurements of the response of any given transmission must be made at such a short distance from its source that the contributions from all other contemporary transmissions are relatively negligible ; but this again involves the question of whether the wave-form of a transmission changes to any appreciable extent with the distance from the source. This matter has been investigated theoretically by R. M. Wilmotte of the National Physical Laboratory, and is discussed in Appendix No. 3. It appears that in general measurements made at distances of the order of one wave-length could be regarded as satisfactory for the specification of the transmission as far as the frequency region in the neighbourhood of the main frequency of the transmission is concerned, but that this would not necessarily be true for harmonics of the main frequency. The

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matter is too complicated for any general statement on this point, and any given case would require separate consideration on the lines indicated by Wilmotte. This aspect of the matter is not, however, of very great importance as far as the legislative side of the question of interference is concerned, for harmonics are not a necessary and inevitable feature of communication in the sense that a main-frequency band is necessary and inevitable, and what is ultimately required in connexion with harmonics is not their specification within stated limits, but their total elimination. Wireless technique has already advanced very far in this direction, and the effective total elimination of harmonics is not too much to expect in the future. In the meantime, the known methods of measurement of signal strength could, if required, be adapted for the purpose of obtaining an approximate measure of the harmonic radiation.

(*b*) *The elimination of the receiving circuit characteristics from the measurements*

It is obviously desirable that the specification of a transmission should be stated in terms of the characteristics of the transmission alone, without reference to those of any receiving circuit with which the necessary measurements had been carried out. The matter is very clearly described in the paper by Chester Snow[4] from which the following quotation is taken :—

> " If the radiation from a given sending radio station is received simultaneously by a set of receiving circuits whose constants are properly chosen, the arrangement constitutes a spectroscope. . . . Or if the message can be repeated, a single circuit may be used and its constants varied. . . . When the circuit possesses very high selectivity, so that it may be considered to respond to a definite frequency only, then if it be tuned successively to various frequencies and the mean square of the current be measured for each, this will be proportional to the spectral distribution of intensity in the radiation. Such an ideal case is not always attainable in practice. The observed current as a function of frequency is often profoundly affected by the character of the receiving circuit. The resonance curve of every circuit possesses a finite width, due to the fact that there is a dissipation of energy in the circuit. A spectral structure in the radiation, which is fine compared to the resonance curve of the circuit, may be entirely masked. The problem which then presents itself is to find how the circuit impresses its character upon the record in order to make corrections for it, and draw inferences as to the character of the radiation which may be, at first glance, entirely hidden in the observations. It is analogous to attempting to resolve a light source by means of a spectrum of insufficient resolving power, where one makes up for lack of resolving power

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by a quantitative study of the energy distribution in the record. . . ."

The necessity for eliminating the characteristics of the receiving systems from the measurements made therewith, was recognised in the original equivalent decrement proposals, where it was suggested that this object could be achieved by the simple subtraction of the decrement of the receiving circuit from the apparent " equivalent decrement " of the transmission as measured with the given receiving circuit. Now it can be shown that in certain ideal simple cases (e.g., trains of exponentially damped oscillations) this procedure has some degree of theoretical justification and should result in approximately the same value for the " equivalent decrement " of the transmission, whatever the decrement of the receiving system on which it is measured. In general, however, this simple process has no theoretical justification whatever, and would in many cases give rise to completely misleading conclusions. As an apparent way out of a difficulty, the proposal has an attractive simplicity, but it cannot be too strongly emphasised that it is in general an entirely arbitrary proceeding and is in no useful sense a solution of the problem.

The mathematical basis of the quantitative study referred to in the above quotation is outlined in Appendix No. 2. It is essentially a matter of finding the solution of a certain type of integral equation. This has been studied by mathematicians, but for the most part the existing purely mathematical work on this subject is of academic rather than of practical value. Some more practical types of solution are given in the paper by Snow,[4] which paper is, in fact, wholly devoted to this aspect of the interference problem. The matter has hardly as yet received the attention it deserves, for it is only in recent years that it has entered the region of applied science. It is to be hoped that further research will be devoted to the matter by mathematicians with a view to arriving at as simple and practicable a solution as may be possible within the limitations of the problem.

In connexion with this same aspect of the matter it should be noted that a further possibility exists, namely, the utilisation of a resonant receiving system of such high selectivity (i.e., low damping) that it can be considered to be virtually a line spectroscope. In other words, a receiving system of such high resolving power that the spectrum of its response can be regarded as a copy of that of the excitation. The construction of a plain inductance and capacity receiving system of this character is impossible, short of using solid mercury coils at the temperature of liquid helium, but there are mechanical resonant systems which can be considered to approximate sufficiently closely to it for practical purposes. Two examples suggest themselves immediately, the quartz crystal, operated piezo-electrically, and a long steel rod in end vibration. A few notes on the experimental aspect of this possibility will be found in Section 10.

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10. The Experimental Measurement of the Mean-power Spectrum of a Transmission

It has been shown in Appendices Nos. 1 and 2 that the measurement of the mean-power spectrum of a transmission requires the measurement of the integral with respect to time of the square of the current induced in a receiving system, the measurement being made at a sufficient number of settings of the tuning adjustment of the receiving system to enable a smooth curve to be drawn through the graphed observations. The time of integration cannot be specified beforehand, but must be such that the value of the integral is proportional to time.

As already stated, there is at present no instrument available for this measurement, and it appears that the practical application of the ideas developed in this Report must await the development of suitable apparatus. In the meantime it will be well to record briefly the negative results of experimental work devoted to this end at the National Physical Laboratory, for the information and guidance of any future workers on this subject.

Two possibilities have been investigated. The first was in effect a balanced air-thermometer method. Two glass bulbs, of about 5 cm. diameter, were connected by a U-tube manometer containing aniline or a light oil. Provision was made for either bulb to be opened to the atmosphere. The bulbs were silvered internally and each was enclosed in a wooden box packed with cotton wool. Into each bulb was sealed a resistance element (5 to 10 ohms) with low resistance sealing wire electrodes brought out through the walls of the bulbs. Fig. 10 will give some idea of the instrument. The idea of the method was that if sufficiently good thermal insulation of the bulbs could be attained the displacement of the liquid column in the manometer tube would be proportional to

$$\int_0^t i^2\,dt,$$

where i is the current passed through one of the resistance elements.

To test the apparatus a constant current was passed through one element and the displacement of the manometer column recorded against time. It was found that the displacement increased fairly rapidly for a few seconds and thereafter remained nearly constant, increasing very slowly indeed. This indicated that after a few seconds an equilibrium was established between the energy supplied and that lost through imperfect thermal insulation. It was considered that the probability of the attainment of sufficiently perfect thermal insulation was not enough to justify the further prosecution of the idea. Moreover, the instrument would not have been very conveniently applicable to the immediate problem, owing to the difficulty of re-establishing a thermal equilibrium between the two bulbs between successive measurements.

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A second method that was tested was the use of a Grassot fluxmeter in association with a thermo-junction. The fluxmeter is essentially a highly damped galvanometer with control reduced to the practicable minimum. The theory of this instrument indicates that its deflection should be proportional to the quantity of electricity passed through it, provided certain conditions are

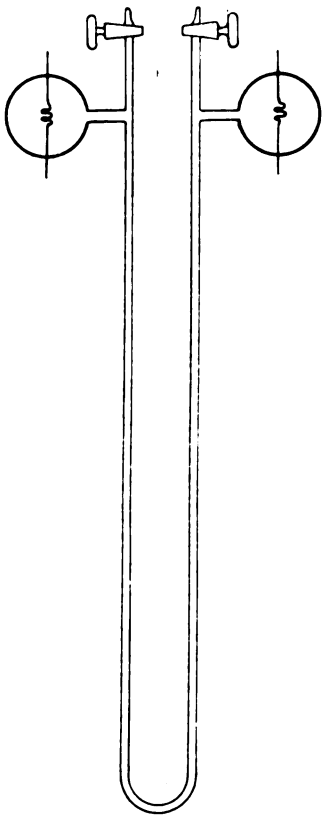

FIG. 10.—*Air-thermometer for measurement of current.*

fulfilled. Its possible applicability to the present problem was tested by passing through the instrument small currents of known amount for various different time intervals up to a total of from 5 to 20 seconds. The actual time intervals were measured very accurately to a tenth of a second by means of a drum recorder. A comparison of coulombs with deflection, however, showed that the deflection per coulomb was by no means constant, varying by 50 per cent. or more in a manner depending on the actual time distribution of the current. A further investigation of the theory of the instrument indicated, as the probable explanation of this

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variation, that the instrument can only be expected to integrate in the manner desired if the moving system comes completely to rest in the time interval between successive impulses. That condition was not fulfilled in the test measurements and would not be fulfilled in a reception measurement. The method was, therefore, abandoned.

A further possibility is the association of a thermo-junction and some form of recording string galvanometer or oscillograph. This, however, would involve the integration of the record by some mechanical means and would be exceedingly laborious when applied to a large number of measurements.

It might be mentioned that a method of a totally different character has also been considered, one which at first sight appears to eliminate the integral equation from the problem. This is the utilisation of the high selectivity attainable with certain mechanical resonant systems, e.g., a long steel rod in end vibration. The idea is, briefly, the transformation of the radio-frequency components of the transmission to the resonant frequency of the mechanical resonant system, which can then be used, in effect, as a spectroscope of such high resolving power that the width of its own frequency band can be neglected.

There are, however, many serious practical difficulties in the realisation of this method, and in addition one theoretical objection which invalidates the idea. Whatever method of heterodyne transformation is employed, using a pure sine wave local oscillation of frequency $\omega/2\pi$, the resonant system will respond to any constituent frequency $\omega^1/2\pi$ of the signal current which satisfies the equation

$$\omega \sim \omega^1 = n,$$

where $n/2\pi$ is the resonant frequency of the resonant system. There are two values of ω^1, namely $(\omega \pm n)$ which satisfy this equation. Thus the spectrum of the transmission is analysed, not with a line, but with two lines, $2n/2\pi$ cycles apart, or rather with two very narrow bands of the same separation. This will not be very serious if n is very small compared with ω, but that condition cannot always be fulfilled.

In addition, the response of the mechanical resonant system will be of a character which is not readily measurable or interpretable. In fact, it is difficult to see how any satisfactory measure of the energy of its oscillation can be obtained without some sort of integrating apparatus of the kind required for the other suggested method.

It should already be clear that the experimental side of the measurement contemplated in this Report presents difficulties towards the solution of which very little progress has yet been made. It is to be hoped, however, that work will continue to be directed to this end as the advance in technique which is required would be of great value in itself, apart from the immediate objective. An integrating instrument of the characteristics described above would certainly find very useful application in many other fields.

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11. Conclusion

It is hoped that the above cursory review will have served to indicate both the importance of the subject of the mutual interference of wireless transmissions and the difficulties which at present lie in the way of its effective control by international regulation. The distinctive feature of the matter, which is at the same time the origin of these difficulties and the reason why they are still unsolved, is that the source of interference is not the periodic and quasi-stationary element of the transmission, but the non-periodic and transient elements. Periodic phenomena are fairly well understood and a practical technique is ready to hand for their quantitative specification. This is the well mapped field of Fourier's theorem. Interference problems, on the other hand, take us into the relatively unexplored field of Fourier's integral. Here both theory and technique are comparatively undeveloped, probably because of the less obvious practical significance of these branches of mathematics and electricity, and because of the greater complexity of the theory and the much greater difficulty of the technique involved. This review will have served a very useful purpose if it stimulates a wider interest in this unexplored region.

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APPENDIX 1

THE FREQUENCY ANALYSIS OF THE INTENSITY OF AN ELECTRIC FIELD

LET $E(t)$ be the instantaneous value of the electric intensity at time t at some point in space. Its variation with time is quite unspecified except that its total integral

$$\int_{-\infty}^{\infty} E(t)\, dt$$

is finite. Subject to this condition, it can be expressed (by Fourier's Integral Theorem) in the form

$$E(t) = \frac{1}{\pi}\int_0^{\infty} d\omega \int_{-\infty}^{x} E(\mu) \cos\omega\,(\mu - t)\, d\mu$$

where μ is a variable of integration.

This holds for all values of t. To demonstrate the physical significance of the right-hand side, it can be written

$$E(t) = \frac{1}{\pi}\int_0^{\infty} (a\, d\omega \cos\omega\, t + b\, d\omega \sin\omega\, t)$$

where

$$a = \int_{-\infty}^{\infty} E(\mu) \cos\omega\ \mu\, d\mu \text{ and } b = \int_{-\infty}^{\infty} E(\mu) \sin\omega\,\mu\, d\mu.$$

The elements $a\, d\omega$ and $b\, d\omega$ define in amplitude and phase a sine wave intensity of frequency $\omega/2\pi$ and lasting for all time. The total intensity is expressed as the sum of an infinite series of such sine wave components, corresponding to a continuous distribution over the whole spectrum.

A form which is more compact and convenient for manipulation can be obtained by utilising the properties of complex or "two dimensional" numbers.

From the above definitions of a and b

$$a - j\, b = \int_{-\infty}^{\infty} E(\mu)\epsilon^{-j\omega\mu}\, d\mu$$

and

$$E(t) = \text{real part of } \frac{1}{\pi}\int_0^{\infty} (a - j\, b)\epsilon^{j\omega t}\, d\omega.$$

Thus the complex number

$$X(\omega) = a - j\ b = \int_{-\infty}^{\infty} E(\mu)\epsilon^{-j\omega\mu}\, d\mu$$

is characteristic of the given intensity variation $E(t)$ and expresses its spectra[1] distribution.

If the intensity is zero or effectively zero for all values of t outside the range t_1 to t_2, then

$$X(\omega) = \int_{-\infty}^{\infty} E(\mu)\epsilon^{-j\omega\mu}\, d\mu = \int_{t}^{t_2} E(\mu)\epsilon^{-j\omega\mu}\, d\mu.$$

(Conversely it can be shown by the theory of contour integration that, subject to the convergence of the integrals concerned, the artificial function

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$f(t)$ which is zero for all values of t outside the range t_1 to t_2 and within this range has the same values as some analytic function $\phi(t)$, can be represented by

$$f(t) = \text{real part of } \frac{1}{\pi} \int_0^\infty \epsilon^{-j\omega t} \, d\omega \int_{t_1}^{t_2} \phi(\mu) \epsilon^{-j\omega\mu} \, d\mu.$$

In other words, the expression on the right has the useful property that it vanishes for all values of t outside the given range and is equal to $\phi(t)$ for all other values of t. This representation has not hitherto received the attention it deserves in relation to the analysis of circuit problems.)

Suppose now that $E(t)$ is zero outside the range t_1 to t_2 and for all other values is given by

$$E(t) = \sum_{n=1}^{n=\infty} (\alpha_n \cos n\,p\,t + \beta_n \sin n\,p\,t)$$

Then

$$X(\omega) = \int_{t_1}^{t_2} \left(\sum_{n=1}^{n=\infty} \alpha_n \cos n\,p\,\mu + \beta_n \sin n\,p\,\mu \right) \epsilon^{-j\omega u} \, d\mu.$$

If the magnitude $| X(\omega) |$ be plotted against ω, it will be found, in general, that the curve will have pronounced peaks in the neighbourhood of those values of ω for which $\omega = np$, $n = 1, 2, 3$, &c. The greater the magnitude of $t_2 - t_1$ (compared with $2\pi/p$) the more pronounced will these peaks become, until ultimately, when $t_2 - t_1$ is very large indeed, the curve will approximate to a discontinuous series of isolated ordinates such that for any given integral value of n

$$\alpha_n^2 + \beta_n^2 \simeq a(\omega)^2 + b(\omega)^2$$

where $a(\omega)$, $b(\omega)$ are the values of a and b corresponding to $\omega = np$. Thus, Fourier's theorem is really a special limiting case of the more general proposition known as Fourier's Integral. Interference problems lie in the region where the time interval $t_2 - t_1$ is not in general sufficiently large compared with the main frequency $p/2\pi$ for the ordinates outside the peaks to be neglected, and the more general integral proposition has, therefore, to be applied instead of the simple special case.

The analysis can be illustrated by reference to the case of n complete sine waves, i.e.,

$$\begin{aligned} E &= \hat{e} \sin pt \text{ for } 0 \leqslant t \leqslant n\mathrm{T} \ (\mathrm{T} = 2\pi/p) \\ &= 0 \text{ for all other values of } t. \end{aligned}$$

$$X(\omega) = \hat{e} \int_0^{n\mathrm{T}} \sin p\mu \epsilon^{-j\omega\mu} \, d\mu$$

$$= \hat{e} \left[\frac{\epsilon^{-j\omega\mu}}{p^2 - \omega^2} (p \cos p\mu + j\omega \sin p\mu) \right]_0^{n\mathrm{T}}$$

$$= 2\hat{e} \frac{j\,p\,\epsilon^{-j\pi n\omega/p}}{p^2 - \omega^2} \sin \frac{\pi n \omega}{p}.$$

It will be more convenient to put $\frac{\omega}{p} = m$, so that $X(mp)$ defines the amplitude and phase of the element having a frequency m times that of the main frequency.

$$X(mp) = \frac{2\hat{e}}{p} \frac{j \sin \pi mn}{1 - m^2} \epsilon^{-jmn\pi}.$$

Therefore

$$| X(mp) | = \frac{2\hat{e}}{p} \frac{\sin \pi mn}{1 - m^2}.$$

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Note that

$$\lim_{m \to 1} | X(mp) | = \frac{2\ell}{p} \lim_{m \to 1} \frac{n\pi}{1+m} \left| \frac{\sin n\pi(1-m)}{n\pi(1-m)} \right| = \frac{n\pi\ell}{p}.$$

Thus

$$\frac{| X(mp) |}{| X(p) |} = \frac{2}{n\pi} \left| \frac{\sin \pi mn}{(1-m^2)} \right|.$$

This ratio, considered as a function of m, has a maximum value 1 when $m = 1$ (i.e., at the main or transmitter frequency, and a series of maxima on either side of this main peak, these subsidiary maxima occurring in the neighbourhood of points for which

$$| \sin \pi mn | = 1$$

$$\text{i.e.,} \quad m = \frac{2r+1}{2n} \qquad r = 1, 2, 3, \text{ \&c.},$$

$$\text{or} \quad \omega = (2r+1)\frac{p}{2n}.$$

The ratio is shown plotted against m for $n = 50$ in Fig. 6. Notice that the peaks occur at frequencies differing by $\frac{1}{nT}$ cycles per second, so that the shorter the number of complete cycles the closer they will be. Notice further that apart from the sine wave fluctuation given by the sine πmn term, the ratio is inversely proportional to n, so that the greater the total number of complete oscillations in the original excitation, the more nearly does it approach a single frequency disturbance, a conclusion which is otherwise obvious. It also shows, however, that a comparatively small number of oscillations will have a very appreciable spectral spread, and is by no means a single frequency disturbance.

We must now consider the relation of the above to the spectral distribution of the energy of any given excitation. For a given field intensity $E(t)$ the instantaneous energy per unit volume of the medium is proportional to $E(t)^2$, and the total energy of the complete excitation will be proportional to

$$\int_{-\infty}^{\infty} E(t)^2 \, dt$$

per unit volume, or if $E(t)$ is zero except between t_1 and t_2, to

$$\int_{t_1}^{t_2} E(t)^2 \, dt.$$

Now

$$E(t) = \text{real part of } \frac{1}{\pi} \int_0^{\infty} (a - j\,b)\epsilon^{j\omega t} \, d\omega$$

where

$$a - j\,b = \int_{t_1}^{t_2} E(\mu)\epsilon^{-j\omega\mu} \, d\mu$$

(i.e., a and b are functions of ω but not of t).

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Multiplying the above equation by $E(t)\,dt$ and integrating from t_1 to t_2 gives

$$\int_{t_1}^{t_2} E(t)^2\,dt = \text{real part of } \frac{1}{\pi}\int_0^{\infty} d\omega.\quad a - jb\int_{t_1}^{t_2} E(t)\epsilon^{\,j\omega t}\,dt,$$

and since

$$\int_{t_1}^{t_2} E(t)\epsilon^{\,j\omega t}\,dt = a + jb$$

$$\int_{t_1}^{t_2} E(t)^2\,dt = \text{real part of}\frac{1}{\pi}\int_0^{\infty}(a^2 + b^2)\,d\omega$$

$$= \frac{1}{\pi}\int_0^{\infty}(a^2 + b^2)\,d\omega = \frac{1}{\pi}\int_0^{\infty} |\,X(\omega)\,|^2\,d\omega.$$

The energy distribution is, therefore, proportional to the square of the ordinates of the $|\,X(\omega)\,|$ curve, and $|\,X(\omega)\,|^2$ may legitimately be termed the energy spectrum of the disturbance. In fact $|\,X(\omega)\,|^2$ is proportional to the energy per unit volume of the medium per unit range of the spectrum. It is a function of frequency and of the shape of the disturbance, but is not a function of time.

It is proposed in this report that $|\,X(\omega)\,|^2$, determined experimentally for an element of any actual transmission of such duration (t_1 to t_2) that

$$|\,X(\omega)\,|^2 \propto t_2 - t_1$$

for any value of ω, is the proper basis for rating the interference producing capability of the transmission.

If the above condition is satisfied, $|\,X(\omega)\,|^2/(t_2 - t_1)$ plotted as a function of frequency, can be described as the mean-power spectrum of the transmission.

It is now necessary to consider how the response of a tunable circuit of known constants depends upon $X(\omega)$ and can be utilised for its determination. This is considered in Appendix No. 2.

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APPENDIX 2

THE RESPONSE OF A RESONANT SYSTEM TO AN IMPULSIVE EXCITATION

CONSIDER a quantity $E(t)$, which is an "artificial" function of time, that is to say, a function which is represented by different analytic functions for various ranges of t. Suppose it to be zero, or negligibly small, for all values of t outside a range t_1 to t_2 and identical in magnitude with some analytic function $\phi(t)$ for all other values of t.

It has been shown that $E(t)$ can be expressed in the form

$$E(t) = \text{real part of } \frac{1}{\pi}\int_0^\infty X(\omega)\epsilon^{j\omega t}\,d\omega,$$

where

$$X(\omega) = \int_{-\infty}^{\infty} E(t)\epsilon^{-j\omega t}\,dt.$$

For purposes of computation, we have

$$\int_{-\infty}^{\infty} E(t)\epsilon^{j\omega t}\,dt = \int_{t_1}^{t_2} \phi(t)\epsilon^{j\omega t}\,dt.$$

This, however, is incidental, and attention will be confined to the integral in terms of the artificial function $E(t)$.

As already explained, this form of representation expresses $E(t)$ as the sum of an infinite number of infinitesimal simply periodic terms $X(\omega)\,d\omega\,\epsilon^{j\omega t}$.

It is important in relation to what follows to realise that the formula from which $X(\omega)$ is computed implies that each of these infinitesimal simply periodic elements is infinite in duration, with its time origin at minus infinity. In other words, the impulsive disturbance is, so to speak, a resultant interference pattern produced by an infinite number of simply periodic terms, each of which began at $-\infty$ in time. Up to the time t_1, the interference between all these elements is complete and results in a zero magnitude. From t_1 to t_2 the resultant is the analytic function $\phi(t)$, and from t_2 onwards the terms annul each other again and give a zero resultant.

The response of a resonant system to such an excitation can now be considered. A circuit consisting of an inductance L, a resistance R and a capacity C in series with a source of electro-motive-force can be taken as a simple example. Representing the e.m.f. as the real part of $(a - jb)\ \epsilon^{j\omega t}$ it is well known that the response of the circuit can be expressed as the sum of two terms. The first is proportional to $\epsilon^{j\omega t}$ and represents a quasi-stationary state of forced oscillation. The second will be proportional to $\epsilon^{(-\alpha + jp)t}$, where α and p are positive numbers depending on L, C and R only. This represents a free oscillation with an exponentially decreasing amplitude. If the origin of time is at minus infinity, the magnitude of this second term will be zero at any current instant t on account of the term $\epsilon^{-\alpha t}$. (It must be remembered that t functions both as a current co-ordinate and a measure of duration. If the origin of t is at $-\infty$, t is infinite at any given instant and $\epsilon^{-\alpha t}=0$.) Thus, if the origin of t is at $-\infty$, the first term alone (i.e., the forced oscillation) will completely represent the response of the circuit at any current instant t. In other words, the free oscillation started so long ago that it has already decreased to zero.

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If, therefore, the origin of t is at minus infinity, the current flowing in the circuit can be completely represented by

$$i = \text{real part of } \frac{(a - jb)\epsilon^{j\omega t}}{R + j(\omega L - 1/\omega C)},$$

$$\text{i.e., } i = \text{real part of } \frac{\omega}{L\{2\omega\alpha + j(\omega^2 - \omega_0^2\}} (a - jb)\epsilon^{j\omega t}$$

where

$$\alpha = R/2L$$

and

$$\omega_0^2 = 1/LC.$$

Whatever the character of the resonant system, the current i will be of the form

$$i = \text{real part of } Y(\omega, \omega_0)(a - jb)\epsilon^{j\omega t}$$

where $Y(\omega, \omega_0)$ is a complex number depending on ω and ω_0, ω_0 being a quantity which can be varied continuously by a suitable continuous variation of one of the constants of the circuit (e.g., C).

With the above introduction it is easy to see that the total response of the circuit to an excitation $E(t)$ of the character described above can be obtained by the summation of elementary responses derived from the elements $X(\omega)\epsilon^{j\omega t} d\omega$, i.e.,

$$i = \text{real part of } \frac{1}{\pi}\int_0^\infty Y(\omega, \omega_0) X(\omega)\epsilon^{j\omega t} d\omega.$$

It is important to realise that this is the complete solution of the second order differential equation involved, comprising in a single expression both the complementary function and the particular integral.

From the present point of view, the desired quantity is not i but the integral energy of the response, i.e.,

$$Q = \int_0^\infty i^2 dt.$$

It is easy to show, exactly as in Appendix 1, that

$$Q = \frac{1}{\pi}\int_0^\infty |Y(\omega, \omega_0)|^2 |X(\omega)|^2 d\omega$$

or, since Q will be a function of ω_0, i.e., will depend on the frequency to which the resonant system is tuned,

$$Q(\omega_0) = \frac{1}{\pi}\int_0^\infty |Y(\omega, \omega_0)|^2 |X(\omega)|^2 d\omega.$$

The proposal outlined in this Report is that $|X(\omega)|^2$, a function of ω depending on the shape of the excitation $E(t)$ is the proper basis for rating the interference producing capabilities of $E(t)$. The experimental measurement involved is the determination of $Q(\omega_0)$ for a sufficient number of values of ω_0 to enable a smooth curve to be drawn exhibiting Q as a function of ω_0. The quantity $|Y(\omega, \omega_0)|^2$ can be regarded as known for any given receiving circuit. It is in any case determinable by laboratory measurements. The final problem is, therefore, the solution of the above integral equation for $|X(\omega)|^2$. It is obviously of the type

$$F(x) = \int_0^\infty G(x, y) f(y) dy$$

where F and G are known and f is to be determined, and is known as an integral equation of the first kind.

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The mathematical solution of this equation is known and is discussed, for example, in Whitaker's *Modern Analysis* and Bôcher's *Introduction to Integral Equations*. Such mathematical solutions are, however, not readily applicable to the present problem, since $F(x)$ is not known as an analytic function, but as a curve of more or less complicated shape which may not be conveniently expressible as an explicit analytic function.

A very good start in this problem has been made by Chester Snow in the paper already referred to, but there is scope for considerable further research in this subject.

To some extent this research waits on a satisfactory solution of the experimental part of the problem, in order that information may be available as to the kind of $F(x)$ curves that will be encountered in practice. No such measurements have yet been made—at least, none have been published. It is true that two experimental curves were given in Chester Snow's paper, but these cannot be regarded as really representative of practical transmission curves. No integrating apparatus was used, the receiving circuit response being measured on a hot wire ammeter. This restricts the type of signal e.m.f. to one that will give a constant reading on such an instrument. Such curves are only of value in affording experimental confirmation of the methods of solution of the integral equation described in the text.

Chester Snow's paper is, nevertheless, a very valuable contribution to this subject, and should be consulted by any who wish for information on the practical aspect of the matter.

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APPENDIX 3

THE EFFECT OF DISTANCE ON THE SPECTRUM OF A TRANSMISSION

By R. M. WILMOTTE, B.A., A.M.I.E.E.

AN antenna system with given surroundings will, when suitably excited, transmit electro-magnetic waves of any desired frequency. The field strength at any point will depend on this frequency and there are many causes which may produce considerable and somewhat complicated variation of field strength with frequency.

As already explained in the body of this Report, a signal of finite duration transmitted from the antenna can be interpreted as the resultant of trains of sine waves of constant amplitude and of infinite duration and, theoretically, of all frequencies, but in practice of all frequencies included in a certain band of finite width centred on the main or nominal frequency of the transmission. Each component will be transmitted in a somewhat different manner from that of every other component of different frequency. The spectrum of the radiation at any given point will, therefore, depend to some extent on the position of that point relative to the antenna.

This fact constitutes a difficulty in obtaining a suitable definition of the interfering capability of the radiation and in devising a method for measuring it. It is the object of this Appendix to consider briefly this aspect of the matter.

If all the effects to be discussed below could be calculated, a knowledge of the spectrum at any point would be sufficient for the calculation of the spectrum at all other points, and the difficulty of finding a suitable expression for the measurement of interference would disappear. In certain simple cases, the calculation can be carried out to a sufficiently close degree of approximation, but in general the work involved would be prohibitively laborious.

The various factors which give rise to a variation of the spectral constitution of the radiation with distance will now be considered separately, in order to obtain some idea of the nature and degree of the variation, and to find whether, in any given case, it may or may not be neglected. The matter can only be outlined here. A more detailed account will be found in the author's paper on the subject, published in the *Journal of the Institution of Electrical Engineers*.[14]

(a) *Proximity Effect*

The existence of a variation in the spectrum with distance from the transmitter can readily be seen from Hertz's equations. The field strength at any point due to the radiation from a current element $i\delta s \sin \omega t$ of length δs, situated at the origin and along the Z axis can be expressed in the usual polar co-ordinates in the form :

$$H_r = H_\theta = E_\phi = 0$$

$$H_\phi = \frac{i\delta s \sin\theta}{cr}\left\{\frac{1}{r}\sin\omega\left(t-\frac{r}{c}\right)+\frac{\omega}{c}\cos\omega\left(t-\frac{r}{c}\right)\right\}$$

$$E_\theta = \frac{i\delta s \sin\theta}{r}\left\{\frac{1}{cr}\sin\omega\left(t-\frac{r}{c}\right)+\omega\left(\frac{1}{c^2}-\frac{1}{\omega^2 r^2}\right)\cos\omega\left(t-\frac{r}{c}\right)\right\}$$

$$E_r = \frac{2i\,\delta s\cos\theta}{r^2}\left\{\frac{1}{c}\sin\omega\left(t-\frac{r}{c}\right)-\frac{1}{\omega r}\cos\omega\left(t-\frac{r}{c}\right)\right\}$$

where c is the velocity of light.

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From the present point of view the interesting fact about these equations is that at a given distance the field depends on the frequency. Now a complex current variation of the type considered in this Report will in effect radiate energy at all frequencies comprised within a certain band of finite width. The field due to any one constituent frequency component will be radiated in accordance with the above equations, and will, therefore, vary with the position of the point considered. The variation with position will, however, be different for each frequency. Thus the spectrum will vary with the position of the point considered. Moreover, since the equations for H_ϕ, E_θ, and E_r are different the spectrum of each of these components will vary with the position of the point.

As the distance is increased the energy of the radiation tends towards a spectral distribution which will be the same for E_θ or H_ϕ. (The component E_r will become negligibly small.) In fact, for distances greater than one wave-length the spectral distribution may be taken as constant with respect to distance. Even then the spectrum of the field will not be the same as that of the current in the antenna. A component differing in frequency from the fundamental by m per cent. will produce a signal strength which will be effectively m per cent. larger.

The above equations refer to a current element only, and in general the total field strength due to the whole antenna would have to be found by an integration process. At distances large compared with the aerial, however, the current element formulæ can be utilised, treating the whole antenna as an element.

It is interesting to note that an effect similar to that described above has been noticed in acoustics. C. F. Sacia [13] mentions that "... a difference in quality of transmitted speech can be detected when the speaker's lips are within 2 cms. of the transmitter diaphragm."

A further factor, intimately related to the one considered above, may be effective in certain cases. The distribution of the current in the aerial depends to some extent upon frequency. For a complex excitation of the kind considered, the spectral constitution of the current in any current element will depend to some extent on the position of that element in the antenna system. This aspect of the matter is described in the next section.

(*b*) *Effect of Directivity*

With few exceptions, unless an antenna extends over a distance comparable with a wave-length, the directive effects are negligible. Small changes of frequency will, in general, produce very small changes of directivity, even in many sharply directive systems, but if the spectrum band extends over a wide range of frequency ($\pm$ 20 per cent., say), very different directive effects may be found between the fundamental and the more distant frequencies.

In sharply directive systems, the various portions of the system are usually so adjusted that the radiations from them add in a given direction. In that case, a small variation of frequency will produce but a second order change in the intensity in that direction. When this condition is not fulfilled, sharply directive systems may produce enormous changes in directive effects for comparatively small changes of frequency. There are, moreover, certain antenna systems, such as the Franklin system, in which part of the current loops are cancelled as far as radiation is concerned. In such cases, a relatively small change in the frequency, may considerably alter the current distribution and produce appreciably different directive effects.

It is evident that any variation of the directive effects with frequency will produce different spectra in different directions. Now the change in directivity may not only be in the horizontal plane but also in the vertical. This may be of importance, for a beam sent upwards will be reflected back to earth, so that, at a distance from the antenna, the spectrum may be very different owing to downcoming waves.

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(*c*) *Effect of Earth*

The resistivity and dielectric constant of the earth have different effects on waves of different frequency. Waves of some frequencies may be absorbed and reflected differently from others.

It is clear from all the preceding considerations that any measurement of the spectral distribution of the radiation of a given transmission, with a view to rating its interfering capability, would require to be supplemented by an investigation of the possible variation of this distribution with distance.

Apart from the present application to the problem of interference, this aspect of the radiation process is of general interest, particularly in relation to transmissions having an appreciable spectral spread.

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APPENDIX 4

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SPECIAL REPORT No. 9

A CRITICAL REVIEW OF LITERATURE ON AMPLIFIERS FOR RADIO RECEPTION

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INTRODUCTORY NOTE

THE present report contains a review and general bibliography of the more recent literature on the design of radio receivers. These have been prepared at the National Physical Laboratory on behalf of the Radio Research Board with the object of indicating the lines along which future research may be profitably undertaken. It is hoped that this survey may be of use to workers who are studying the properties of radio receivers. To bring out the relative importance of the different sections the review is essentially critical and all the papers cited have been examined from this point of view. It is believed that the bibliography is practically complete from 1916 to April, 1929, but it is realised that some of the literature on the subject has possibly been overlooked. For the early literature prior to 1916 citation is made only of those publications which are of historical value or otherwise of outstanding importance.

The review is divided into four main sections :—I. Radio Frequency Amplifiers ; II. Rectification ; III. Audio Frequency Amplifiers ; IV. Measurements. No mention has been made of either the antenna system or the loud speaker or telephones, since these subjects open up a large field of literature which is not relevant to amplifier design.

The main sections of the review have been sub-divided and given index marks, e.g., 1.1, 1.2, 1.3, etc., and a critical essay based on the study of the literature of each sub-section is included. References within the text refer to the bibliographical sections and abstracts following the sub-section review.

The sub-section bibliographies are further sub-divided and indexed as 1.1.1, 1.1.2, 1.1.3, etc. References are given throughout in the following manner—title, author, journal, year, volume, pages, the volume being expressed in clarendon type. The abstracts given have been made by reference to the original paper in every

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case. The references in each division of the sub-sections are consecutively numbered and arranged in chronological sequence. Cross references are given where a paper has been inserted in more than one section owing to its treatment of several subjects, and an abstract of that part of the paper relative to the particular section is given in each section in which it appears. Any critical comments expressing the compiler's views on a paper are given at the end of the abstract and are inset to distinguish them from the purely abstracted portion.

A list of journals which have been consulted follows this introductory note. Where possible the abbreviations of the titles of the journals have been taken from the World List of Scientific Periodicals. A main index will be found at the end of the report containing the abstract number and the page reference of each abstract cited.

Sections I, III and IV have been prepared by Mr. H. A. Thomas, M.Sc., and Section II by Mr. F. M. Colebrook, B.Sc. The compilers are indebted to Dr. R. L. Smith-Rose for much helpful criticism and to Mr. C. W. Spencer for his work in collecting material and preparing abstracts.

H. B. JACKSON,
Chairman,
RADIO RESEARCH BOARD.

August, 1929.

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LIST OF PERIODICALS CONSULTED

(Abbreviations according to World List of Scientific Periodicals where possible)

Abbreviated Title.	Full Title.
Ann. Physik.	*Annalen der Physik.*
Ann, Post. Télégr. Téléph. ..	*Annales des Postes, Télégraphes et Téléphones.*
Arch. Elektrotech.	*Archiv. für Elektrotechnik.*
Bell. System Techn. J. ..	*Bell System Technical Journal.*
Bull. Soc. franç. Élect.	*Bulletin de la Société français des Électricians.*
Bull. Soc. franç. Radio ..	*Bulletin de la Société français Radio.*
C.R. Acad. Sci. Paris	*Comptes Rendus Hebdomadaires des Séances de l'Acadèmie des Sciences.*
Elect. J.	*The Electric Journal.*
Elect. Rev. Lond.	*The Electrical Review, London.*
Electrician	*The Electrician.*
Elect. Wld.	*Electrical World.*
Elek. Nachrichten	*Elektrische Nachrichten-technik.*
Elektrat. Laborat. Tôkyô Japan Researches	*Researches of the Electrotechnical Laboratory, Tokyo, Japan.*
Elektrotech. u. Maschinenb. ..	*Elektrotechnik und Maschinenbau (Die Radiotechnik).*
Elektrotech. Z.	*Elektrotechnische Zeitscrift.*
Elettrotechnica	*L' Ettrotechnica.*
Experimental Wireless ..	*Experimental Wireless and the Wireless Engineer.*
Gen. Elect. Rev...	*General Electrical Review.*
J. Amer. Inst. Elect. Eng. ..	*The Journal of the American Institute of Electrical Engineers.*
J. Franklin Inst.	*The Journal of the Franklin Institute.*
J. Inst. Elect. Eng. Jap. ..	*The Journal of the Institute of Electrical Engineers of Japan.*
J. Instn. Elect. Eng.	*The Journal of the Institution of Electrical Engineers.*
J. Inst. Wireless Tech... ..	*The Journal of the Institute of Wireless Technology.*
J. Opt. Soc. Amer.	*Journal of the Optical Society of America and Review of Scientific Instruments.*
J. Post Office Elect. Eng. ..	*The Post Office Electrical Engineers' Journal.*
Jahrb. drahtl. Telegr.	*Jahrbuch der Drahtlosen Telegraphie und Telephonie. Zeitschrift für Hochfrequenztechnik.*
Lumière Élect.	*La Lumière Électrique.*
Onde Élect.	*L'onde Électrique.*

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Phil. Mag.	*The London, Edinburgh and Dublin Philosophical Magazine and Journal of Science.*
Phil. Trans. Roy. Soc. ..	*Philosophical Transactions of the Royal Society of London.*
Phys. Rev.	*The Physical Review.*
Phys. Z.	*Physikalische Zeitschrift.*
Proc. Inst. Radio Eng... ..	*Proceedings of the Institute of Radio Engineers.*
Proc. Nat. Acad. Sci. Wash. ..	*Proceedings of the National Academy of Sciences, Washington.*
Proc. Phys. Soc. Lond. ..	*Proceedings of the Physical Society of London.*
Proc. Radio Club Amer. ..	*Proceedings of the Radio Club of America.*
Proc. Roy. Soc.	*Proceedings of the Royal Society of London.*
Q. S. T.	*American Q. S. T.*
Q. S. T. Franç. Radioélect. ..	*Q. S. T. Français et Radioélectricité.*
R. Accad. Navale, Livorno, Instit. Elettrotechn. e. Radiotelegrafico	*Accademia Reale Navale di Livorno, Instituzione Elettrotechnica e Radiotelegrafico.*
Radioélectricité	*Radioélectricité.*
Radio Eng.	*Radio Engineering.*
Radio f. Alle	*Radio für Alle.*
Radio News	*Radio News.*
Radio Rev.	*Radio Review.*
Rep. Brit. Ass.	*Report of the British Association for the Advancement of Science.*
Rev. Gen. Elect...	*Revue Générale de l'Électricité.*
Sci. Pap. Bur. Stand. Wash. ..	*Scientific Papers of the U.S. Bureau of Standards, Washington.*
Telefunkenztg.	*Telefunkenzeitung.*
Telegr. u. Fernspr. Tech. ..	*Telegraphen und Fernsprech Technik.*
Tijdschr. ned. Radiogenoot. ..	*Tijdschrift van het Nederlandsch Radiogenootschap.*
Tôhuku University Technical Reports	*Tôhuku University Technical Reports*
T.S.F. Moderne	*T.S.F. Moderne.*
Wireless Age	*Wireless Age.*
Wireless Wld.	*Wireless World and Radio Review.*
Z. f. Phys.	*Zeitschrift für Physik.*
Z. tech. Phys.	*Zeitschrift für technische Physik.*

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SECTION I.—RADIO-FREQUENCY AMPLIFIERS

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1.1. CIRCUIT ARRANGEMENTS AND GENERAL PROPERTIES

THE bibliography of this section is arranged in chronological order, thereby demonstrating the growth of circuit design during the past ten years. The simplest circuit arrangements will not be discussed as they are dealt with in the theoretical portions of the Review.

1.1.1. Dual or "Reflex" Amplification Circuits

The first departure from the simplest circuits was the Dual or "Reflex" Amplification Circuit.

The aim in all types of circuit adopting the reflex principle is to make one valve serve a double purpose, usually as a radio and audio-frequency amplifying stage simultaneously. The rectifier may be a valve or crystal but is more usually the latter, serious difficulties arising with a valve rectifier due to retro-action effects. In order to make one valve serve the double function of radio and audio-frequency amplifier, it is necessary to shunt the secondary of the audio-frequency transformer so that the radio-frequency input circuit has a low impedance. The presence of this by-pass condenser produces not only a reduction in the audio-frequency amplification of this stage but also gives a bad audio-frequency amplification characteristic. For an efficient radio-frequency stage, a large shunting condenser is necessary, whereas the reverse is required for an efficient audio frequency stage.

The design of necessity resolves itself into a compromise, and the values of this condenser suggested by different investigators (see Abs. 1.1.1/3, 6) reveal the marked differences of opinion current on the subject. Values lying between zero and 2,000 micro-microfarads are suggested. In the former case, the self-capacity of the transformer winding is considered sufficient effectively to by-pass the radio-frequency energy, while in the latter case 2,000

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$\mu\mu$F. is required. In view of recent work on inter-valve transformers, it is known that such a shunt capacity would completely ruin the linear frequency response characteristic of the audio-frequency stage. The difficulty of maintaining the high audio-frequencies so necessary for good articulation is so great that a suggestion has been advanced that no attempt should be made to preserve frequencies above 5,000 cycles per second, and that the by-pass condenser should be arranged to preserve the lower frequencies as best it can.

There is another source of distortion in reflexed circuits due to the serious retro-action effects between the radio and audio-frequency sections of the system. Since these overlap so seriously, the grid potential excursions are of a complex nature and radio-frequency transients (such as atmospheric noises) give rise to complex audio-frequency disturbances.

The reflex principle has been extended to four and five stages in which the properties of the circuits become very involved and considerable difficulty is experienced in maintaining stability. The reflex circuit was produced at a time when sheer amplification was the only demand made upon a receiver, but as ideas have changed and greater refinement of acoustic fidelity in reproduction is demanded, the difficulties arising from this economy in valves have been found to be so serious that the system is now in disfavour for telephony reception. For continuous-wave work it possesses economical advantages, but here the demands of greater selectivity in radio-frequency amplifiers tend to overrule its doubtful advantages. The tendency throughout the development of receiver design has been to designate each stage to a particular function and design it for that particular purpose. From this point of view, the reflex circuit is an example of bad design.

1.1.2. Super-Regenerative Circuits

The next step was the introduction of " triggered " circuits in which a weak signal was permitted to throw an unstable system into self-oscillation. By this means, a critical adjustment could be obtained and the weak signal E.M.F. would produce a large resultant output due to the unstable system having been upset and thrown into oscillation. The difficult problem with all such circuits was the method of restoring the oscillating system to its original " threshold " condition of stability.

The most important practical circuit of this type is undoubtedly the Armstrong Super-Regenerative Receiver.

The operation consists in permitting oscillations of a free nature to be built up independent of the value of the impressed E.M.F., thus permitting a very weak input to trip an oscillating system. The Turner relay was the first attempt to make use of this principle. In this case, the grid was negatively biassed so that the system was

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only just stable. A small input would carry the potential of the grid over the threshold value and a free oscillation would start building up to the limiting value of the valve. The system was returned to its initial sensitive state by means of a mechanical relay operated by the increase in the anode current of the valve, this relay short circuiting the feed-back coil and permitting the potential of the grid to drop back below the threshold value. The clumsiness of this mechanical device limited the arrangement to the reception of weak Morse signals and special calling devices. BOLITHO (see Abs. 1.1.2/3) improved the mechanical relay of TURNER by connecting a second valve to the oscillating circuit with a reversed retro-action connection. When the threshold of the first valve was overcome and a free oscillation started in the system, the reversed retro-action damped out the oscillation and permitted the grid of the first valve to return below the threshold value. This Bolitho circuit represents the second step in the development of the regenerative principle.

The third step, due to ARMSTRONG (see Abs. 1.1.2/1), is based on the fact that if a periodic variation is introduced into the system making the resistance of the input circuit alternatively positive and negative, free oscillations proportional to the amplitude of the impressed E.M.F. will occur during the negative resistance period. For maximum amplification, the starting and stopping of the oscillations should be at as low a frequency as possible, and herein lies the chief difficulty in practice. Either the grid or the anode of the receiving valve may be oscillated periodically by coupling to the audio frequency oscillator. The first oscillator may be used for rectification as well, but the problem of cascade amplification with these systems is difficult, unless filter circuits are adopted. A low periodic audio-frequency is necessary to obtain maximum amplification, since the period during which the system remains unstable determines the magnitude of the built up self-oscillations ; but this periodic effect produces a constant audible hum in the receiving telephones (called the " whistle "), which although permissible for the reception of continuous waves is most objectionable for the reception of telephony.

The main defects of the average commercial super-regenerative receiver are :—

(*a*) A definite limit is set to signal strength and no useful system of audio frequency magnification has been evolved.

(*b*) The valve is rectifying extremely inefficiently.

(*c*) The " whistle " must be rather too distinctly audible if good signals are to be obtained on medium wave-lengths.

(*d*) The characteristic re-heterodyning effect brings in a great deal of heterodyne noise.

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(*e*) Since the effective resistance of the input circuit is periodically made zero, the circuit is rendered oscillatory, and therefore radiates. It cannot therefore take the form of an open antenna due to interference with other receivers. A small frame antenna is necessary and this partially detracts from the natural sensitivity of the receiver, owing to the small signal E.M.F. induced in it. This defect can be overcome by suitable arrangements, notably a radio-frequency amplifying stage between the frame and the first rectifier (see Abs. 1.1.3/11).

(*f*) The complexity of the circuit behaviour gives rise to transient phenomena which produce serious distortion effects when the receiver is used for the reception of telephony. Special filtering arrangements are necessary to overcome this defect.

A further development of the regenerative principle has been made by FLEWELLING (see Abs. 1.1.2/4), in which a simplification of the circuit arrangement is claimed by making use of the drop of potential across condensers to stimulate self-oscillation whilst the necessary damping and quenching effect is effected by a leak resistance. It is claimed that audio-frequency amplification can be added, but the " whistle " difficulty remains and the circuit is very noisy.

1.1.3. THE SUPERSONIC-HETERODYNE RECEIVER

The next important development is the circuit known as the Supersonic-Heterodyne Receiver.

The received radio-frequency signal is heterodyned by a local oscillator, the beat frequency being usually about 200 kilo-cycles per sec. Owing to the ease with which low radio-frequency amplifiers can be designed, it is possible to use two or three stages in cascade without special stabilising devices. This " intermediate frequency " amplifier effectively avoids the difficulty of amplifying at the reception frequency. The circuit principle is fundamentally that of a frequency changer. The possible received frequency is only defined by the lower limit of the local heterodyne oscillator, and since this may be as high as we choose, the receiver is suitable for very high frequencies.

In some cases the rectifier is made to play a double role, namely, that of detector and local heterodyne oscillator, and in this case, or with separate heterodyne, the input circuit is oscillating and must, therefore, be sensibly non-radiating, hence the almost universal use of a small frame. This limitation on oscillatory receivers in general is a severe one and seriously weakens their advantages.

Since the object of the receiver is to eliminate the difficulties inherent in any system of amplification at high radio-frequencies, it is clear that its chief application lies in the reception of these

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high radio-frequencies, and although the receiver can be used for frequencies approaching that of the fixed frequency intermediate amplifier, there appears to be little merit in so using it, since the first rectifier will appreciably diminish the overall amplification and it would be better to use normal amplification methods for such low radio-frequencies.

The difficulties of amplification at high radio-frequencies are so. great that the supersonic-heterodyne receiver has almost a monopoly for this type of reception.

The frequency transformation produces distortion, and the receiver is not very satisfactory for telephony reception unless special filtering methods are adopted.

The chief practical difficulties which have had to be overcome are as follows :—

(*a*) The sensitivity is somewhat dependent upon the amplitude of the beat oscillation, this latter being largely dependent upon the frequency of the beat oscillator. The sensitivity wave-length characteristic is therefore far from uniform, and it is found that the receiver has good properties over only a small working range of wave-lengths, unless the beat oscillator is capable of being variably coupled to the input circuit.

(*b*) The main tuning is somewhat dependent upon the heterodyne tuning and strength, due to coupling effects, and exact calibration of either the main tuning condenser or oscillator condenser is not usually possible.

(*c*) The intermediate-frequency amplifier is liable to pick up powerful stations working on or near its natural frequency, and since this is definitely fixed, this source of interference can only be eliminated by elaborate screening or suitable filters.

(*d*) The first rectifier is working very inefficiently owing to the self-oscillating behaviour superimposed upon the rectification action.

For pure continuous-wave reception, however, the selectivity can be made very great by using a very selective multi-stage intermediate-frequency amplifier. The use of iron cored radio-frequency transformers has somewhat improved this selectivity, and if the output is re-heterodyned to an audio-frequency and passed on to a selective audio-frequency tuned amplifier, or through a band pass filter, the resultant selectivity may be made as great as we choose.

In view of the narrow wave-length band allocated to most American stations, greater selectivity is required than in this country and the supersonic-heterodyne receiver is still very popular in the United States, whereas requirements of acoustic fidelity have been more eagerly sought in Great Britain, and this receiver

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has been largely replaced by types less selective but giving purer telephony reproduction.

The use of double grid valves for frequency changing has been suggested by four French experimenters (see Abs. 1.1.3/4). In this case one grid is fed with the signal E.M.F., while the other is supplied with the local heterodyne. Another minor modification is to use an intermediate-frequency amplifier working with the sum of the two frequencies instead of the difference (see Abs. 1.1.3/5). It is claimed that since the intermediate-frequency amplifier is working at such a high frequency, little extraneous interference from powerful stations is likely, but the difficulties of using such a short wave amplifier nullify the advantages of frequency changing.

1.1.4. "Neutrodyne" Circuits

The next stage in development is the use of neutralisation—that is, the attempt to balance the inter-electrode capacity effects by a bridge network so that an efficient stable stage can be used at wave-lengths where it would normally oscillate, due to capacitative coupling between the input and output circuits via the anode-grid capacity. The most popular form of this circuit is the Neutrodyne Circuit.

The advantages of the neutrodyne circuit are that by balancing the valve capacities by an external bridge network, the maximum amplification due to an efficient tuned system in the anode circuit of the radio-frequency stage can be obtained with stability, whereas, prior to this, stability was obtained only by potentiometer controlled grid damping with its attendant disadvantages :—

(*a*) A reduction in stage amplification.

(*b*) The insertion of a load upon the input tuned system.

(*c*) Distortion due to grid current.

Great improvements in radio frequency amplification have been made due to the use of such neutralised circuits in conjunction with more efficient tuned output circuits. Greater selectivity and amplification with stability at all wave-lengths can thus be obtained.

The neutralising condenser balances the anode-grid capacity, and assuming this capacity to be pure, it is theoretically possible to obtain stability at all wave-lengths. In practice, the anode-grid impedance is not a pure capacity, partly due to the imperfections in the valve but chiefly due to the circuit arrangement. In this case, balancing would require an impedance in place of the condenser, and this impedance would need to be variable in the same manner as the anode-grid impedance varied with frequency. Since this is impossible of attainment, the neutralised stage is found in practice to be stable over a limited range only, and usually at the highest frequency to which the receiver is designed to operate,

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instability occurs, or if not actual self-oscillation, considerable retro-action.

The amplification-frequency response characteristic is not uniform and it appears theoretically impossible completely to nullify the effects of inter-electrode capacity.

In general therefore the neutralised circuit offers an excellent solution to the problem of undistorted radio-frequency amplification at medium and high frequencies, but the solution is not perfect and there is much room for improvement.

1.1.5 Reaction Control Devices and the "Reinartz" Circuit

Departing temporarily from the question of amplification, we notice the growth of certain retro-action control devices which have improved the ease of manipulation considerably. The most significant of these circuits is the Reinartz Circuit.

Here, a condenser is so arranged between the anode and grid that critical retro-action can be obtained without seriously disturbing the tuning of the main primary circuit. Its chief claim is that the adjustment of critical retro-action is easier than with the older method of inductively coupling the anode and grid circuits. Care must be taken to screen this condenser adequately to prevent "hand effects," and the principle has been useful when embodied into neutralised circuits where it can raise the amplification appreciably. However, all such retro-action devices suffer from the general indictments against retro-action controls of any kind.

1.1.6. Resistance-Capacity-Coupled Radio-Frequency Amplifiers

The desire to obtain a radio-frequency amplifier which is capable of operation over a limited though extensive wave-length range without tuning controls has led to improvements in Resistance-Capacity-Coupled Radio-Frequency Amplifiers.

The use of resistance-capacity-coupled amplifiers for frequencies above 300 kilo-cycles per sec. (1,000 metres) has usually been considered inadmissible, but if great care is taken, such aperiodic radio-frequency amplifiers can be constructed for frequencies up to 1,500 kilo-cycles per sec. (200 metres). The difficulties arise from stray capacity effects due to wiring, valve holders and the inherent residual capacities in the "pinch" of the valve. By using valves in which the electrode leads are not bunched in a cap, eliminating the holder, and spacing short leads to the valve, it is possible to reduce these shunt capacity effects, which largely determine the effective impedance of the anode resistance.

Such attempts are naturally limited by the definite characteristics of the valves, and have not been found to be very successful. The

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rapid reduction in amplication as the frequency is raised due to inter-electrode capacity effects is inevitable.

An attempt to reduce stray capacity effects to a minimum has been made by LOEWE, who inserts two valve electrode systems with the necessary coupling units inside one large glass bulb. By this means, the leads from one valve to the other are reduced to a minimum and shunt capacity effects are thereby made small, but severe coupling effects are present due to the proximity of instability, which detracts somewhat from the obvious advantages of the arrangement—simplicity and portability.

Two types have been manufactured (see Abs. 1.1.6/3), a radio-frequency unit, and a rectifier and audio-frequency unit. The former consists of two stages complete with resistances, leaks and coupling condensers mounted in one bulb, and suitable for frequencies up to 1,000 kilo-cycles per sec. (300 metres). The latter bulb contains a rectifier, one resistance-capacity-coupled amplifying stage and a power stage. This latter stage is only suitable for medium power outputs and the audio-frequency type is in general applicable to small outputs only. The complete 5 valve receiver made by the use of the two Loewe valves is portable and light.

1.1.7. METHODS OF REDUCING INTER-STAGE COUPLING

We now come to a consideration of the improvements that have been made within the last few years and which have so materially improved the performance and reliability of the radio receiver. These improvements to a great extent have consisted in reducing stray couplings, preventing retro-active effects and obtaining high-amplification stages with stability. Screening (see Abs. 1.1.7/1) has been developed and coils and complete stages are now screened from their neighbours. Retro-action due to high tension battery resistance has been reduced by the use of shunted anode feed resistances, and care has been taken to reduce coupling due to wiring by the use of special concentric and twin leads (see Abs. 1.1.7/2).

1.1.8. CIRCUITS USING MULTIPLE-ELECTRODE VALVES

Several forms of valves have been made with four electrodes, but the one which perhaps is the best known and of the greatest present interest, is that containing a cathode, two grids and an anode. The uses and types of two-grid valves may be divided into three general classes depending upon the functions of the two grids; and to a considerable extent the use to which the valve is to be put influences the details of the design. These classes include the "screen-grid" valve, the "space-charge-grid" or "double-grid" valve, and various "double function" and miscellaneous valves.

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The "double-grid" valve, or "space-charge-grid" valve, takes its name from the fact that the inner grid is used in such a way as to reduce the space charge around the cathode and thereby the internal anode resistance. This is done without any decrease in amplification factor, hence there is an increase in mutual conductance corresponding to the decrease in anode resistance. The inner grid or space-charge-grid, in its simplest connection, is kept at a fixed positive potential with respect to the filament. The outer grid is the control electrode and performs exactly the same function as in the triode. The arrangement of the two grids is therefore just opposite to that of the screen-grid valve. Owing to its properties it can be used to reduce the voltage of the high-tension battery necessary for a given amplification.

The earliest work on the screen-grid valve was done by W. SCHOTTKY followed by that of HULL (see Abs. 1.1.8/9), who provided more complete shielding. The inner grid is the control electrode while the outer grid acts as a screen. (see also 1.2.8).

Its high voltage factor, together with an anode-grid capacity of the order of 0·1 $\mu\mu$F., gives at once a valve in which retro-action and consequent instability are almost eliminated. The internal resistance of the valve is very high, usually between 500,000 and 1,000,000 ohms, and consequently the selectivity is greater than that obtained with the triode. The valve is very useful in most types of radio-frequency amplifiers and lends itself particularly to supersonic-heterodyne intermediate-frequency work where stability can be obtained when several tuned systems have their condensers mechanically connected to a common tuning control.

The modern tendencies in design can be sectionalised into :—

(*a*) Improved screening of the receiver as a whole from the input circuit, and the screening of individual stages and coils.

(*b*) Simplicity of control by the use of mechanically connected condensers, fixed filament resistors and adjustable capacity retro-action.

(*c*) Selectivity and stability with high amplification by the use of neutralised circuits or tetrodes.

(*d*) High amplification and selectivity by the improvement in radio frequency coil and transformer design.

(*e*) Care in inter-stage wiring to minimise capacitative and inductive coupling, and the use of special wiring systems.

(*f*) The prevention of retro-active coupling due to battery resistance, by anode-feed systems and radio-frequency chokes.

The American tendency has been towards simplified gang control, screening and neutralisation with 5, 6 or even 7 stages.

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In this country the tendency has been rather similar, neutralised transformer-coupled receivers of not more than 3 stages becoming more popular. Care in wiring lay-out and the use of anode-feed resistances to prevent battery coupling are becoming more general. It is worthy of notice that the modern tendency is towards the perfection of the " straight " receiver, i.e., where each stage fulfils one, and only one particular function for which it can be satisfactorily designed. Screen-grid valves or neutralised triode systems are displacing regenerative and supersonic-heterodyne receivers chiefly due to the greater refinement demanded in fidelity of reproduction.

1.1.9. Filter Circuits

A filter circuit or system is an arrangement of electrical circuits which possesses definite frequency characteristics. One type, called a "high pass" filter, produces very little attenuation of frequencies above a selected value, while heavily attenuating all frequencies below that value, while the other type, called a " low pass " filter, will attenuate all frequencies below a definite selected value. By a combination of these two types, a " band pass " filter can be obtained which will only pass frequencies lying between two definite values. Such filters are used to eliminate interference at both radio and audio-frequencies.

A filter circuit consists of a series of similar circuits linked by conductive, capacitative, inductive or mutual inductive coupling. Theoretically, an infinite number of "links" are required to produce a very high ratio of attenuations between the frequencies " passed " to those " stopped " and give a sharply marked " cut-off," i.e., a very steep curve of frequency—attenuation in the boundary region between " pass " and " stop," but the theoretical requirements can be closely approximated by a finite and frequently small number of links.

The selective properties of a wave-filter are peculiar to the steady state, and become sharply defined only as the steady state is approached. To this fact is to be ascribed the failure of wave-filters to suppress irregular and transient interference, such as atmospherics, in anything like the degree with which they discriminate against steady-state currents outside the transmission range. This limitation is common to all types of selective networks, and restricts the amount of protection it is possible to secure from transient or irregular interference. This is due to the fact that the the transients contain frequencies in the " pass " region and these are, necessarily, " passed."

Owing to the mathematical nature of the subject, a closer analysis is not possible in essay form and the reader is referred to the Bibliography on the subject for details of design in any particular case.

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Improvements are described for broadcast wave-lengths, the secondary of the audio-frequency transformer being shunted by a condenser of 50 micro-microfarads. The crystal rectifier is still maintained.

3. **Double Magnification Circuits :** W. James : *Wireless Wld.*, 1923 **12**, 164-67 and 211-14.

The author advocates the use of a shunt capacity of 2,000 micro-microfarads across the audio-frequency intervalve transformer. A crystal rectifier is adopted and a complete four-valve receiver is described in which it is claimed that the performance is equivalent to that of six normal valves.

The shunting effect of the radio-frequency by-pass condenser across the primary of the inter-valve transformer will produce not only a reduction in the amplification of this stage, but a very bad audio-frequency response characteristic. The value of the by-pass condenser suitable for radio-frequency amplification is most unsuitable for audio-frequency amplication if a reasonably linear audio-frequency response characteristic is required.

4. **Faithful Reproduction in Radio-Telephony :** L. C. Pocock : *J. Instn. Elect. Eng.*, 1924, **62**, 791-807.

(See Abs. 3.4/4.)

5. **Discussion of a Double Magnification Circuit :** A. Cazes : *Onde Élect.*, 1926, **5**, 425-28.

It is shown that the output can be materially increased by permitting the first stages of an amplifier to function in several ways. The employment of a frequency changer can make each valve work on three different frequencies.

Although possibly satisfactory from the point of view of pure amplification, the distorting possibilities of such an arrangement if used as a receiver for telephony are so numerous that its utility is severely limited.

6. **By-Pass Condensers in Reflex Circuits :** D. Kingsbury : *Wireless Wld.*, 1926, **19**, 546-50.

The difficulty is discussed of maintaining the high audio-frequencies so necessary for good articulation and the suggestion is made that the by-pass condensers across the audio-frequency transformer should be reduced to a minimum and in certain cases dispensed with altogether, it being shown that the self-capacity of the windings and the input capacity of the valves may sometimes be sufficient. Since a compromise is necessary in any dual circuit, the suggestion is put forward that the audio-frequency transformer

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should have a rising characteristic up to 5,000 cycles and then a cut-off above this frequency. This means that reception should be limited to 5,000 cycles on either side of the carrier wave.

The paper clearly shows the difficulties attendant on the use of the dual circuit and indicates why it has now been abandoned in all cases where purity of reproduction is of importance.

7. **The Performance of Reflexed Valves :** D. KINGSBURY : *Experimental Wireless*, 1927, **4**, 547-51.

The author describes a phenomenon which he observed with reflexed circuits, consisting of a howl produced due to a special form of interference between the radio-frequency and the audio-frequency impulses on the first grid. The phenomenon and its cause are described in some detail, and the author concludes with a plea for more serious attention to the possibilities of this type of circuit.

1.1.2. SUPER-REGENERATIVE CIRCUITS

1. **Super-Regenerative Valve Circuits :** E. ARMSTRONG : *Proc. Inst. Radio. Eng.*, 1922, **10**, 244-68.

A system of circuits is described whereby the effective resistance of a regenerative circuit is periodically made positive and negative, but predominantly positive. Such a circuit responds to impressed electromotive forces due to incoming signals, and free oscillations are set up during the negative resistance epoch, the final amplitude being proportional to the exciting E.M.F. The amplitude increases with time so long as the effective resistance is negative. The forced oscillation under these conditions is comparatively small. This increase in amplitude of the free oscillation is termed super-regeneration, and the phenomenon is applied to the reception of radio-telegraphy and telephony, for which circuits are given. The means required to carry the circuits through the necessary resistance variations are also given.

2. **Super-Regeneration :** P. DAVID : *Onde Élect.*, 1923, **2**, 222-29.

This paper describes, with details of inductances, capacities, etc., a receiver of the Armstrong type specially suitable for reception of the [illegible] kilo-cycles per sec. (450 metres) wave sent out by the Ecole Supérieure des P.T.T. and also gives instructions for the adjustment of the set.

3. **The Bolitho Circuit :** J. B. BOLITHO : *Wireless Wld.*, 1923, **12**, [illegible].

With reference to the principles utilised in the Armstrong super-regenerative circuit, a claim to priority has been advanced by J. B. BOLITHO. The present paper gives an account of the Bolitho circuit extracted from the patent specification. The circuit consists of a valve oscillator in which the oscillations take place intermittently, this being effected by a generator in the anode circuit at a frequency much lower than that of the received signal. A simplified wiring diagram is given and the action of the set explained.

4. **A New Circuit :** : *Wireless Wld.*, 1923, **12**, 74.

The Flewelling circuit. This is claimed to be a simpler arrangement than the Armstrong super-regenerative circuit. It makes use of the drop of potential across condensers to stimulate self-oscillation, whilst damping is provided by a leak resistance. Audio frequency amplification can be added, but the circuit is noisy and the whistle cannot be eliminated.

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5. **On Super-Regeneration :** E. O. HULBURT : *Proc. Inst. Radio Eng.*, 1923, **11**, 391-94.

Two simple super-regenerative circuits are described with the values of the electrical constants.

(See also Abs. 1.2.4/7.)

1.1.3. THE SUPERSONIC-HETERODYNE RECEIVER

1. **A New System of Short Wave Amplification :** E. H. ARMSTRONG : *Proc. Inst. Radio Eng.*, 1921, **9**, 3-27.

(See Abs. 1.2.4/2.)

2. **The Supersonic-Heterodyne—its origin, development and some recent improvements :** E. H. ARMSTRONG : *Proc. Inst. Radio Eng.*, 1924, **12**, 539-52.

The paper describes the development of the supersonic-heterodyne receiver from a war-time invention into a commercial form of broadcast receiver.

The success of the development is due to a reduction in the number of valves used, this being obtained by combining the oscillator and rectifier into one valve, reflexing, self-heterodyning and improving the design of the radio-frequency transformer. The arrangement is not so satisfactory as when each valve has an independent function, but is essential for portable requirements.

3. **The Origin of the Supersonic-Heterodyne :** W. SCHOTTKY : *Elek. Nachrichten*, 1925, **2**, 454-56.

In this paper the author traces the origin and development of reception methods in Germany corresponding to Armstrong's supersonic-heterodyne.

(See also Abs. 1.1.3/7.)

4. **Use of Double-grid Valves in Frequency Changing Circuits :** J. DE MARE, R. BARTHÉLEMY, H. DE BELLESCIZE and L. LEVY : *Onde Élect.*, 1926, **5**, 150-80.

(See Abs. 1.2.4/11, 2.3.3/13, 14, 15, 16.)

5. **The Infradyne :** H. GREEN : *Radio News*, 1926, **8**, 356-57.

A description is given of a circuit differing from the supersonic-heterodyne in that, instead of making use of the difference in frequency between the local oscillations and the incoming signal and amplifying this beat, the sum of the frequencies is utilised. Amplification is thus carried out at the high radio frequency of 3,200 kilo-cycles per sec. (95 metres) instead of between 30 and 100 kilo-cycles per sec. (3,000 to 10,000 metres), as in the supersonic-heterodyne. In addition to the total absence of interference from long wave transmitting stations, the advantage is claimed of the complete elimination of harmonics which are so prevalent in supersonic-heterodyne receivers.

The chief merit of the supersonic-heterodyne receiver, namely, the removal of amplification at very high radio-frequencies, has been lost in this circuit. In this system the problem of designing an efficient short wave amplifier has been made more difficult and its rather dubious advantages do not compensate for the difficulties of the short wave amplifier as compared with the intermediate-frequency amplifier of ordinary supersonic receiver practice.

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6. **Supersonic Transformers :** N. W. McLachlan : *Wireless Wld.*, 1926, **19**, 631-34, 680-84, 715-18, 777-78 ; *Wireless Wld.*, 1927, **20**, 21-4.

(See Abs. 1.3/7.)

7. **On the Origin of the Super-Heterodyne Method :** W. Schottky : *Proc. Inst. Radio Eng.*, 1926, **14**, 695-98.

A general historical essay on the origin and development of the supersonic-heterodyne type of receiver.

(See also Abs. 1.1.3/3.)

8. **Super-heterodyne Receivers :** E. H. Ullrich and A. H. Reeves : *Experimental Wireless*, 1927, **4**, 652-56.

When the signal input circuit and the beating oscillator circuit of a supersonic-heterodyne receiver are coupled to the grid of the first detector in any of the usual ways, the tunings of the two circuits are not independent of each other, since there is a certain degree of coupling between them. This not only causes inconvenience owing to the multiple tuning points thus involved, but with the degree of coupling usual for frequencies above 3,000 kilo-cycles per sec. (100 metres), an actual discontinuity appears.

(See also Abs. 1.2.4/15.)

9. **Improvements in Tuned Amplifiers :** Blanchard : *Onde Élect.*, 1927, **6**, 57-70.

(See Abs. 1.1.9/15.)

10. **The Screened Valve in Audio-Frequency Circuits :** N. W. McLachlan : *Wireless Wld.*, 1927, **21**, 536-38, 577-79.

(See Abs. 3.2.7/2.)

11. **The Preliminary Input Valve for Frequency-change Receivers :** O. Knies : *Radio f. Alle.*, 1928, 503-5.

The author points out that such a preliminary radio-frequency stage, in front of the main super-heterodyne receiver, not only increases the sensitivity, but also prevents re-radiation, which is not entirely avoided by the use of a frame aerial. One of the best methods of coupling to the first stage of the receiver is by a tuned transformer, the auxiliary adjustments being a variable condenser and the grid-bias potentiometer. Other methods are illustrated and discussed.

12. **New Receiving Circuits with New Multiple Valves :** M. von Ardenne : *Radio f. Alle.*, 1928, 488-99.

The paper describes a new type of valve suitable for the intermediate frequency amplifier of a supersonic-heterodyne receiver.

(See also Abs. 1.1.8/15.)

13. **A Double Super-Heterodyne : A Description of a Receiver Built by the Author :** J. F. Ramsay : *Experimental Wireless*, 1928, **5**, 669-72.

The author begins by discussing why the super-heterodyne receiver is not more popular in England ; a bad reputation regarding quality may be one cause—this is the fault of manufacturers who strain to get the maximum gain per stage ; another cause is that the admonitions to " cultivate the local station " are obeyed. The double super-heterodyne described and illustrated, when receiving short waves of 45 metres, transforms to 500 metres and amplifies, then transforms again to 2,000 metres and amplifies

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again. When used in this manner ten valves and two oscillators are employed. The broadcast from Schenectady can be made to overload the loud-speaker. No screening has been found necessary.

A very general non-analytical paper.

14. **Recent Developments in Super-heterodyne Receivers :** G. L. BEERS and W. L. CARLSON : *Proc. Inst. Radio Eng.*, 1929, **17**, 501-15.

Major electrical elements of a modern supersonic-heterodyne receiver are briefly discussed in the light of recent developments. A practical automatic volume control is described whereby fading can be almost eliminated. Curves illustrating the overall performance characteristics of the receiver are shown.

The method of measurement, however, is not given.

15. **The Problem of the most Economic Frequency changing in Reception :** F. AIGNER : *Jahrb. drahtl. Telegr.*, 1929, **33**, 9-15.

(See Abs. 1.2.4/23.)

1.1.4. "NEUTRODYNE" CIRCUITS

1. **Resistance Amplifiers :** L. BRILLOUIN : *Onde Élect.*, 1922, **1**, 1-17, 101-23.

The second part of this paper examines the part played by capacity "compensation" in the performance of radio-frequency resistance-capacity-coupled amplifiers.

(See also Abs. 1.4.1/2.)

2. **The Neutrodyne :** W. JAMES : *Wireless Wld.*, 1923, **12**, 67-71.

The paper gives a simple discussion of the theory of the neutrodyne receiver and details for the construction of such a receiver are appended.

3. **Tuned Radio-Frequency Amplification with Neutralisation of Capacity Coupling :** L. A. HAZELTINE : *Proc. Radio Club Amer.*, 1923, **2**, No. 8.

This paper gives a description of the early commercial form of Hazeltine's "neutrodyne" receiver.

This is Hazeltine's original paper on the neutrodyne receiver.

(See also Abs. 1.4.2/6.)

4. **The American Neutrodyne :** K. H. STARK : *Wireless Wld.*, 1925, **16**, 525–27.

A general description of the neutrodyne circuit is given, together with a history of its American development.

5. **New Neutrodyne Arrangements :** F. DUFRENOY : *Onde Élect.*, 1926, **5**, 487-89.

A simple non-mathematical description of some unusual types of neutrodyne circuits is given.

6. **Shielded Neutrodyne Receiver :** J. F. DREYER and R. H. MANSON : *Proc. Inst. Radio Eng.*, 1926, **14**, 217-47, 395-412.

After outlining the development of the neutrodyne receiver, in which capacity coupling is neutralised by the use of suitably arranged coils and condensers, a description, with full details, is given of the construction of

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a neutrodyne receiver in which each stage is screened from the others by enclosing it completely in a copper screen. The advantages of this screening are enumerated.

7. **Discussion on "The Shielded Neutrodyne Receiver," by Dreyer and Manson :** L. A. HAZELTINE : *Proc. Inst. Radio Eng.*, 1926, **14**, 395-412.

(See Abs. 1.1.4/6.)

8. **Tuned Radio Frequency Amplifiers :** R. S. GLASGOW : *J. Amer. Inst. Elect. Eng.*, 1928, **47**, 327-31.

The paper concludes with a discussion of various types of neutralising devices to render a receiver stable over a large range of frequencies.

(See also Abs. 1.2.2/20.)

1.1.5. REACTION CONTROL DEVICES AND THE "REINARTZ" CIRCUIT

1. **The Improved Reinartz Circuit :** J. L. REINARTZ : *Q.S.T.*, 1922, March.

2. **A New Valve Circuit and its Use for the Construction of a Universal Receiver, and a Note on Direct Current Amplification for Automatic Reception :** G. LEITHAUSER : *Jahrb. drahtl. Telegr.*, 1923, **21**, 30-32.

Describes a circuit stated to allow fine regulation of retro-action. This is illustrated and described in detail for use with one and two valves.

3. **Two Reflex Reinartz Circuits :** H. C. EXEL : *Wireless Wld.*, 1924, **15**, 247-48.

In one case a crystal detector and a reflex Reinartz circuit is described, and in the other a valve rectifier is used.

4. **A New Method of Retro-action Control and its Application to a Universal Receiver :** E. FROMY : *Onde Élect.*, 1925, **4**, 149-54.

An ingenious method of controlling the retro-action in a retro-acting valve receiver is described. The oscillating circuit has fixed inductive retro-active coupling, such as the Hartley circuit, and the tendency to oscillate is controlled by varying the effective resistance of the circuit. This is effected by connecting a special condenser across the tuned circuit. It consists of a double condenser, controlled by one knob, and is so arranged that as the capacity of one half is increased, that of the other half is decreased by the same amount. One half is connected directly across the inductance, while the other half is connected through a fixed damping resistance. The result is that, as the knob is turned, the total effective capacity remains constant while the effective resistance is continuously and smoothly varied ; this controls the tendency of the receiver to oscillate. Variation of tuning is accomplished by means of a separate ordinary condenser also connected across the inductance. A diagram is given showing how the principle may be embodied in a two valve tuned-anode receiver.

In effect, it is a special form of neutrodyne arrangement in which the unbalancing of the neutrodyne condenser is used as a reaction control. No theoretical considerations are attempted however, and the circuit merely stands as a useful novelty.

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1.1.6. Resistance-Capacity-Coupled Radio-Frequency Amplifiers

1. **Resistance Amplifiers :** S. Brillouin : *Onde Élect.*, 1922, **1**, 1-17, 101-23.

(See Abs. 1.4.1/2.)

2. **Resistance Amplifiers for Short Waves :** G. Beauvais and L. Brillouin : *Onde Elect.*, 1923, **2**, 267-74.

It has frequently been stated that resistance amplifiers are of little or no use for frequencies above 333 kilo-cycles per sec. (1,000 metres). In the present paper experiments are described bearing on this point. It appears that the above statement is correct unless special precautions are taken to eliminate stray capacities. The valves themselves must be at least 10 cm. apart, and if the valves are of the French R-type, special socket fittings must be used to reduce to a minimum capacity effects between the fittings themselves. By reducing stray capacities in these ways and paying attention to the arrangement of the apparatus, it was possible to construct an amplifier of the resistance type which gave good amplification at a frequency of 860 kilo-cycles per sec. (350 metres), using ordinary R-valves. With some valves good results were obtained at a frequency of 1,500 kilo-cycles per sec. (200 metres). Wiring details and diagrams are given.

3. **Loewe Resistance Capacity Amplification Unit :** S. Loewe and M. v Ardenne : *Jahrb. drahtl. Telegr.*, 1926, **27**, 19-20.

The authors describe a new form of amplifying unit, consisting of two valves and the necessary coupling resistances, condensers and grid leaks, all mounted inside a single evacuated bulb. By this means, the various connecting leads are reduced to the minimum possible dimensions so that such a resistance-capacity amplifier can deal satisfactorily with medium wave-lengths. In order to allow of the use of a normal value of high tension voltage along with high anode resistances, the authors make use of tetrodes instead of the usual triodes. An extra potential of about 12 volts is applied to the second grid. There are two types of this unit, one primarily for radio frequencies and the other for audio-frequency amplification.

The undistorted output available from the audio-frequency type is not sufficient to operate a modern loud-speaker, but may be used to supply the grid of a power valve, and the radio-frequency type compares favourably with the more conventional resistance-capacity radio-frequency amplifier using separate components.

(See also Abs. 3.1.1/4.)

4. **Aperiodic High Frequency Amplification :** H. Kröncke : *Wireless Wld.*, 1927, **20**, 27-28.

A brief description is given of the possible wave-lengths at which resistance-capacity coupling can be used, together with a discussion on the merits of the Loewe valve.

1.1.7. Methods of Reducing Inter-Stage Coupling

1. **The Shielding of Electric and Magnetic Fields :** J. H. Morecroft and A. Turner : *Proc. Inst. Radio Eng.*, 1925, **13**, 477-505.

(See Abs. 1.2.9/4.)

2. **Scientific Wiring :** W. B. Medlam : *Wireless Wld.*, 1928, **22**, 449-52.

The author gives theoretical reasons for the adoption of certain wiring layouts in which the use of twin flex and concentric conductors is suggested.

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1.1.8. Circuits using Multiple-Electrode Valves

(a) *The Double-Grid Valve*

1. **The Negatron. A New Negative Resistance Device for Use in Wireless Telegraphy :** J. Scott-Taggart : *Rep. Brit. Ass.*, 1921, **89**, 436, No. 14, and *Electrician*, 1921, **87**, 386.

A thermionic valve is arranged with two flat anodes on each side of the filament. Each anode is connected through an anode battery to the filament so that the electrons emitted by the filament are distributed equally between the two anodes. A control electrode, e.g., a flat grid, is also arranged within the valve between the filament and one of the anodes. This latter anode is called the " diversion anode," while the first one is called the " main anode." By suitably arranging the relationship between the electron emission and the anode voltages, it is possible to obtain a saturation effect. When the main anode voltage is increased, the grid potential is increased ; electrons are diverted from the main anode and the main anode current decreases giving a negative resistance effect.

The main use of the device is as a generator of continuous oscillations for the transmission or reception of continuous waves. It may be used to reduce the positive resistance of a circuit and so produce retro-action effects.

The effect is critically dependent upon the filament current ; if this is too small, or too large, no effect takes place.

2. **The Four-Electrode Valve :** H. Nozières and P. Giroud : *Onde Élect.*, 1924, **3**, 583-90.

The paper deals with the characteristics and behaviour of the double-grid valve, its properties being well illustrated by characteristic curves.

The reduction in battery voltages required is apparent and two circuit arrangements utilising the valve as an amplifier are given.

(See also Abs. 2.3.3/3.)

3. **Negative Resistance Phenomena in Double-Grid Valves :** P. Amye : *Onde Élect.*, 1925, **4**, 297-305.

A number of schemes of connection are given making use of the double-grid valve (a similar valve to Hull's dynatron) in amplification and rectifier circuits. The author concludes that such valve circuits would offer advantages for the reception of distant stations under bad local conditions.

(See also Abs. 2.3.3/7.)

4. **Use of Double-Grid Valves in Frequency Changing Circuits :** J. de Mare, R. Barthélemy, H. de Bellescize and L. Levy : *Onde Élect.*, 1926, **5**, 150-80.

(See Abs. 1.2.4/11, 2.3.3/13, 14, 15, 16.)

5. **Employment of Four-Electrode Valves :** B. Decaux : *Onde Élect.*, 1927, **6**, 1-18.

(See Abs. 1.2.8/5, 2.3.3/19.)

6. **Four-Electrode Valves. Some Applications :** B. Decaux : *Bull Soc. franç. Elect.*, 1928, **8**, 126-30, and *Onde Élect.*, 1928, **7**, 119-24.

An account of the research of the French Military Radio Laboratories into the use of the double-grid valve. They have been used in retro-action and amplifier circuits, advantage having been taken of the possibility of using very low anode voltages to enclose a complete radio-frequency amplifying stage, including batteries, in a screened case.

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(b) The Screen-Grid Valve.

7. **The Marconi Four-electrode Valve and its Circuit :** H. DE A. DONISTHORPE : *Proc. Inst. Radio Eng.*, 1924, **12**, 411-21.

(See Abs. 2.3.3/4.)

8. **Four-electrode Receiving Valves :** A. C. BARTLETT : *Wireless Wld.*, 1925, **17**, 773-74.

(See Abs. 2.3.3/8.)

9. **Characteristics of Shielded-Grid Pliotrons :** A. W. HULL and N. H. WILLIAMS : *Phys. Rev.*, 1926, **27**, 432-38.

The internal feed-back from anode to grid of a triode is prevented by electrically shielding the one from the other. Specially designed valves, thus shielded, are described. The capacity between the anode and control grid was found to be only 0·006 $\mu\mu$ F. in the best models. These valves, therefore, may be expected to behave as strictly unidirectional devices at all frequencies, and this was found to be true. Anode-resistance characteristics are given, and it is shown that the internal anode conductance is negligible compared with the conductance of practical circuits. The anode current thus depends on the grid voltage only, i.e., in place of the seven parameters of the triode there is but one in this tetrode, viz., mutual conductance.

10. **Four-electrode Portable Receiver :** F. L. DEVEREUX : *Wireless Wld.*, 1926, **18**, 544-48.

These are constructional articles describing complete receiving sets using double grid valves operating in the ordinary manner but with low anode battery voltage.

No detailed analysis of operation is given.

11. **Screened Valves :** P. W. WILLANS : *Wireless Wld.*, 1927, **21**, 107-10.

The elementary principle of the screen-grid valve is given, the valves of HULL being described. A general circuit layout of a four-valve receiver is described in detail.

12. **The Cossor Screened Valve :** *Wireless Wld.*, 1927, **21**, 817-19.

A description of construction and characteristic curves.

13. **A Relay-Amplifier for Short Waves :** J. VIVIÉ : *Q.S.T. Franç. Radioélect.*, 1928, No. 56, 21-4.

The circuit consists of a combination of a screen-grid radio-frequency amplifier valve and the " Schnell " triode circuit with electrostatic retro-action coupling. The wave-length range considered is from 100 to 10 metres.

14. **The Five-Electrode Valve for Multiple Functions :** R. BARTHÉLEMY : *T.S.F. Moderne*, 1928, **9**, 396-400.

A description of a circuit (" trisodyne ") in which the three-grid valve gives radio-frequency amplification and a frequency circuit, with retro-action in the primary circuit and also in the beat-note circuit. A receiver using this circuit with four valves (one for medium radio-frequency, one for rectification and one for audio-frequency) gives results " comparable, if not superior," to those obtained with seven or eight valves with the usual frequency-changing devices.

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15. **New Receiving Circuits with New Multiple Valves :** M. v. ARDENNE : *Radio f. Alle.*, 1928, , 488-99.

The paper describes a new type of valve designed by the author in conjunction with LOEWE. It is claimed that a high voltage factor is obtained together with a reduced inter-electrode capacity, making it possible to use six such radio-frequency stages in cascade without trouble. Special care in wiring and screening is necessary and the amplifier may be used as an intermediate-frequency amplifier for short wave reception.

16. **Some Characteristics and Applications of Four-Electrode Tubes :** J. C. WARNER : *Proc. Inst. Radio Eng.*, 1928, **16**, 424-46.

Four-electrode valves may be classified by their designs and uses as "screen-grid," "space-charge-grid" and "double-function" valves. In the screen-grid valve, the inner grid is the control electrode and the outer or screen-grid is kept at a fixed potential. The capacity between the anode and the control-grid is thereby reduced to an almost negligible value. A second result of the screen-grid is a large increase in the amplification factor and the anode-filament resistance without reduction of the mutual conductance. This permits high amplification in connection with high impedance coupling circuits, without undesired retro-action or self-oscillation. The screen-grid principle has been applied to transmitting valves as well as tc receiving valves. Characteristics of these valves are given in detail. In the space-charge-grid valve, the outer grid is the control electrode and the inner grid is maintained at a fixed potential. The purpose of the inner grid is to reduce the effect of the space charge around the filament and thereby reduce the anode resistance of the valve. The space-charge-grid valve performs the same functions as ordinary triodes, but in general has a higher mutual conductance than a triode of similar design. Several double function valves and circuits are described in which both grids act as control electrodes or in which one grid acts as the control-electrode and the other as a combination space-charge-grid and control or output electrode.

These circuits are sometimes useful but are subject to certain definite limitations.

17. **The Screen-Grid Tube :** N. H. WILLIAMS : *Proc. Inst. Radio Eng.*, 1928, **16**, 840-43.

In the screen-grid valve, retro-action due to the anode-grid capacity is reduced to a negligible amount and the anode current is very nearly independent of the anode voltage over the working range. Under these conditions, the voltage amplification becomes the product of the mutual conductance and the load impedance. A high impedance in the anode circuit is obtained by using a low resistance resonance circuit. With good screening, such a circuit may be used without producing self-oscillation. At a frequency of 700 kilo-cycles per sec. (430 metres) a stage amplification of 80 can be obtained.

This is a very simple descriptive paper only.

1.1.9. FILTER CIRCUITS

1. **Physical Theory of the Electric Wave-Filter :** J. A. CAMPBELL : *Bell System Techn. J.*, 1922, **1**, 1-32.

This paper represents an elementary physical explanation of the wave-filter as a device for separating sinusoidal electrical currents of different frequencies. Fundamental theories concerning the energy flow through artificial lines are dealt with, following which wave-filter response curves for certain special cases are given. It is shown that a wave-filter approximates

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to a resistanceless artificial line which is capable of only two fundamental distinct motional states. In one state, the disturbance is attenuated along the line, and there is no flow of energy other than a back-and-forth surge, the intensity of which rapidly falls off along the line. In the other state, there is a free flow of energy without loss from section to section along the line, with no surge of energy between symmetrical sections. The former state gives a stop-band filter, having a definite frequency range, while the latter gives a pass-band filter. Since the attenuation factors vary in geometric progression with the number of sections, a high degree of discrimination between different frequencies may be obtained.

2. **Theory and Design of Uniform and Composite Electric Wave-Filters :** O. J. ZOBEL : *Bell System Techn. J.*, 1923, **2**, 1-46.

This paper gives systematic methods for designing wave-filters together with representative designs. A general theory of design is first considered, followed by an application to the design of a low-pass and a band-pass filter, and lastly, design formulæ are obtained for all the practical uniform wave-filters likely to be encountered in practice.

3. **Electric Filter Circuits :** L. COHEN : *J. Franklin Inst.*, 1923, **195**, 641-54.

The author points out that the calculation of the performance of a filter circuit, from the standpoint of an artificial line, does not bring out clearly the improvement in band-frequency tuning as the number of sections is gradually increased. In the present paper, the system is treated as a coupled circuit, and it is shown that band tuning is obtained only when an infinite number of independent circuits is used, but that the condition can be closely approximated to when only a limited number of sections is employed. From the theory given, curves can be plotted for any individual case, showing the deviation from the uniform band tuning for any limited number of circuits.

4. **Transient Oscillations in Electric Wave-Filters :** J. R. CARSON and O. J. ZOBEL : *Bell System Techn. J.*, 1923, **2**, 1-52.

Previous papers have discussed the steady-state properties of electric wave-filters. The present paper gives the results of a fairly extensive study of their behaviour in the transient state. It appears from this study that the remarkable selective characteristics of wave-filters are peculiarly properties of the steady state, and become sharply defined only as the steady state is approached. To this fact is to be ascribed the uniform failure of wave-filters to suppress irregular and transient interference such as atmospherics, in anything like the degree with which they discriminate against steady-state currents outside the transmission range. This limitation is common to all types of selective networks, and restricts the amount of protection it is possible to secure from transient and irregular interference.

5. **Theory of Electric Filters :** C. LANGE : *Ann. Post. Télégr. Téléph.*, 1923, **12**, 1252-92.

Uniform electric wave-filters of the Campbell type are here treated as a particular case of uniform artificial line composed of recurrent identical " T " or " TT " networks, corresponding to mid-series or mid-shunt termination. Fundamental formulæ in hyperbolic form are established by Kennelly's method for the sending-end impedance, the current ratio and the voltage ratio, in terms of the series impedances, the shunt admittances, the number of sections and the terminating impedance. A general analysis is made as to the conditions to be fulfilled for an artificial line to act as an electric filter, neglecting resistance. The usual 2 and 3-element Campbell filters are treated for the ideal case. If it is desired to take resistance into account, it is recommended to compute the equivalent series impedance and shunt

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admittance for each desired frequency and hence compute the effect of the filter in the circuit by applying the fundamental formulæ for artificial lines.

6. **Wave Filters :** P. LE CORBEILLER and C. LANGE : *Onde Élect.*, 1923, **2**, 560-70.

A theoretical treatment is given of the arrangement of conductors (including inductances, capacities, resistances positive or negative, etc.) in the form of the letter T. KENNELLY has shown that a network composed of inductances, capacities, etc., can, for a given frequency, be replaced by a certain well-defined T for all operations carried out at its terminals. In the present paper a general formula is obtained relative to any T whatever, and by applying this to a line formed of a large number of T's in series, simple conditions are obtained which determine the values of the frequencies which such a line will allow to pass, that is, the frequencies for which the system acts as a filter.

7. **Transmission Characteristics of Electric Wave-Filters :** O. J. ZOBEL : *Bell System Techn. J.*, 1924, **3**, 567-620.

The characteristics of electric wave-filters and systematic methods of deriving them have been considered in previous papers (see Abs. above). The present paper deals with a simple and rapid method of calculating the steady-state transmission losses of wave-filter networks over both the transmitting and the attenuating frequency bands, including the effects of dissipation and wave-filter terminal conditions. General formulæ for any dissymmetrical linear transducer are derived in terms of new parameters, called image parameters. The formulæ lend themselves to graphical representation, and a detailed illustration is given of this chart method of calculation as applied to composition wave-filters.

8. **Theory of Short Filter Circuits :** H. BACKHAUS : *Jahrb. drahtl. Telegr.*, 1924, **24**, 11-16, 39-45, 53-65.

The theory of two or three linked circuits is given, together with methods for determining the shape of the response characteristic. Two and three circuit systems are designed from theoretical considerations, and the experimental results obtained with these networks are shown to agree fairly well with the theory.

9. **Theory of Short Filter Circuits. Part II :** H. BACKHAUS : *Jahrb. drahtl Telegr.*, 1925, **25**, 93-94, 134-38.

A continuation of the previous paper. General formulæ are derived for a four unit system, and experimental resonance curves are given for this case.

10. **Wave Filters :** K. S. JOHNSON and T. E. SHEA : *Bell System Techn. J.*, 1925, **4**, 52-111.

In the first part of this paper the authors deal with the general principles of wave filter design for networks without mutual inductance, while the second part deals with filters using mutual inductance. Impedance and propagation characteristics are fully dealt with, equivalent sections considered, and appropriate formulæ derived.

11. **Production of Single Sideband for Transatlantic Radio Telephony :** R. A. HEISING : *Proc. Inst. Radio Eng.*, 1925, **13**, 291-312.

Part of this paper describes the filtering arrangements adopted to utilise a single side band only at the transmitter. Transmission curves for the filters are given as well as some amplitude-frequency performance curves of the complete transmitter.

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12. **Theory of Electric Filters :** P. DAVID : *Onde Élect.*, 1926, **5**, 5-27, 72-80.

The author shows that the design of filters is easy if the ohmic resistance of the elements is neglected, and each element is represented by its impedance. Different elements are considered and it is shown that, by combining different elementary types to form a composite filter, greater efficiency is obtained than with simple homogeneous filters. The effects due to resistance are then considered, together with the case of filters which cannot be represented accurately by their reactive impedances only. Finally, the advantages to be obtained by introducing mutual inductances into filter circuits are briefly outlined.

13. **Resonance Curves of Electric Filters :** H. GOERING : *Arch. Elektrotech.*, 1926, **17**, 316-36.

In this mathematical paper the behaviour of filter circuits is examined by means of a system of resonance curves. Both symmetrical and non-symmetrical systems are considered, and in both cases resonance curves are obtained for purely ohmic end resistance and for any end impedance whatever:

14. **Filtering Properties of Chains of Electric Circuits :** V. GORI : *Elettrotechnica*, 1927, **14**, 753-57.

This paper is a summary of the work which has been done upon wave-filters. The general problems of a chain of circuits is considered, followed by the selective or filtering properties of these chains. Whilst such filters can give only small protection against disturbances, yet they constitute the best selective circuit against perturbations interfering with the signals.

15. **Improvements in Tuned Amplifiers :** BLANCHARD : *Onde Élect.*, 1927, **6**, 57-70.

Two circuit arrangements are described in which the amplification is practically constant over a small range of frequencies and falls off very rapidly outside this range.

These arrangements, although not filters, have similar properties and are very useful as intermediate-frequency transformers for supersonic-heterodyne receivers.

16. **The Theory of Filter Chains :** H. WINTER-GÜNTHER : *Jahrb. drahtl. Telegr.*, 1928, **32**, 41-46.

The theoretical method of treatment which considers a filter as a coupled oscillating circuit cannot readily be applied to obtain the natural frequencies except in certain limiting cases. The author shows how the use of the system of normal co-ordinates simplifies and extends the application of the method.

17. **The Reidel Wave Stopper :** H. BOCK : *Radio f. Alle.*, 1928, , 541-45.

The author confirms Reidel's original specification and investigates the circuit mathematically, after which he considers some of the practical difficulties, e.g., the difficulty of eliminating harmonics of the local station.

18. **The Filter Problem :** P. OLINET : *Q.S.T. Franç.*, 1928, No. 49, 37-45, No. 51, 3-10.

The article deals with the question of the filtering necessary to make rectified A.C. current from mains suitable for the supply to transmitters from the point of view of quality. The problem is examined in considerable detail by means of curves and simple mathematics. Lastly, single wave filter circuits are examined as an adjunct to the other filters. The question of the anode feed, both for transmission and reception, is examined in detail.

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19. **Wave-Filters :** H. Bock : *Radio f. Alle.*, 1928, , 157-62, 196-200.

The paper treats the subject mathematically.

20. **Fundamental Proposition in the Theory of Electric Filters :** J. B. Pomey : *Rev. Gen. Elect.*, 1928, **23**, 347-50.

The paper gives a solution of the problem of electric filters in terms of Lagrange's equations with generalised variables.

The paper is entirely of a theoretical nature and examines the problem very thoroughly.

21. **Proportions of Symmetrical, Loss-Free Chain Filters between Ohmic Resistances :** R. Feldtkeller : *Elek. Nachrichten*, 1928, **5**, 145-49.

The author criticises the work of Zobel, which, he states, does not permit a clear picture of the variation in damping for the different types of filter. Methods are developed for symmetrical loss-free filters.

22. **Chains of Resonant Circuits :** E. Mallett : *J. Instn. Elect. Eng.*, 1928, **66**, 968-74.

The chain of resonant circuits coupled by mutual inductance is reduced to the more usual type of filter circuit or chain of T-links in Section I. In Section II, a solution is obtained for the currents flowing, and this is applied to the chain of resonant circuits in Section III, and to the case of a chain of asymmetrical T-links in Section IV. In the last Section a graphical construction of the solution is given, and an example is worked out in an Appendix.

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1.2. ANALYSIS OF RADIO-FREQUENCY AMPLIFICATION

1.2.1. General Theory and Transients

Any signalling wave may be resolved into sustained sinusoidal components, which may be thought of as traversing the system as individual currents and recombining at the receiving end to form the reproduced signal. The possibility of such a resolution has been demonstrated mathematically and the formulæ for evaluating the amplitudes and phases of the components are well known. A periodic wave may be expressed as a Fourier series, that is, as the sum of an infinite series of components, the frequencies of which may be thought of as harmonics of a fundamental frequency which is equal to the frequency of repetition of the wave. Such a resolution, however, is not directly applicable to non-periodic or transient waves. A receiver must be capable of receiving any individual symbol regardless of what precedes or follows it. We may, however, resolve such a " transient " by the mathematical device of assuming it to be one cycle of a periodic wave in which the interval between successive occurrences of the disturbance in question approaches infinity. The frequency of repetition is then infinitesimal. The fundamental frequency of the Fourier series and the frequency interval between adjacent components are also infinitesimal ; that is, the series of discrete lines of the Fourier series spectrum merge into a continuous spectrum (see Abs. 1.2.1/12).

For any modulated wave, it is necessary to know the distribution of the energy in the form of a spectrum, in order to find the limits of frequency over which the receiver is desired to operate. Such knowledge is also needed for the design of a " bandpass " filter in the receiving circuits.

Periodic phenomena do not completely represent the operation of a receiver, which is constantly being operated by complex potential waves of non-periodic form. The response of a given amplification system to an aperiodic or quasi-periodic signal is a problem of great complexity as it requires the solution of the response of a given system to a transient impulse. From a practical standpoint, a knowledge of the steady state phenomena is sufficient in the great majority of problems, but in many cases the " transient " behaviour is the dominating factor.

The problem of analysing the behaviour of any electrical system when impressed with a " transient " E.M.F. resolves itself into formulating an expression which gives the current as a function of time if the original E.M.F is similarly expressed as a time function.

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This problem can be solved by the application of Fourier's analysis (see Abs. 1.2.1/23), but is very laborious; a better method is to use "Heaviside's Expansion Theorem" (see Abs. 1.2.1/4, 10, and 15).

1.2.2. The Valve Network

The early work of Richardson (see Abs. 1.2.1/1) and Langmuir (see Abs. 1.2.1/2) gives a general knowledge of thermionics and the characteristics of the triode. Barkhausen (see Abs. 1.2.2/7) and Van der Bijl (see Abs. 1.2.2/5) extended this fundamental work to the examination of the space current obtainable with various designs of electrodes, and Vallauri (see Abs. 1.2.2/3) made use of the characteristics of the triode to examine its properties as an amplifier and rectifier. This work represents the first serious attempt to analyse the properties of the simple valve network. It is based on the characteristics of the triode and assumes that these are linear over the working range. A more complete examination of the single stage case was made by Carson in 1919 (see Abs. 1.2.2/8). Although inter-electrode capacity effects are neglected, the voltage factor and internal resistance of the valve are assumed constant and the effects of grid current are not considered, this work nevertheless represents a definite stage of progress in amplifier theory. The anode current variations are calculated for resistance loads in the anode circuit for known grid potentials. The theoretical problems involved in heterodyne reception are also considered.

It was not until the work of Miller (see Abs. 1.2.2/10) and Ballantine (see Abs. 1.2.7/4) upon the input impedance of a valve network (1.2.2) was completed that the important effects of this impedance began to be realised. All previous theoretical work must therefore be regarded as incomplete and as offering an elementary solution of the problem, roughly applicable under certain cases.

The treatment of the valve circuit as a network and the use of equivalent circuits to represent its properties was first undertaken by Nichols (see Abs. 1.2.2/9), who realised the importance of input impedance effects, including the nature of the output load.

A careful mathematical and experimental analysis of the nature of the output for a single stage of radio frequency amplification applied with a pure sinusoidal input was made by Frayne (see Abs. 1.2.2/12), in which the magnitude of the harmonics produced by the curvature of the characteristics was determined. The frequency used, however, was too low to permit any effects due to the valve capacities taking place, and the results are deducible from the characteristics of the valve.

Llewellyn stated in 1926 (see Abs. 1.2.2/17) that the earlier work summarised in Carson's paper of 1919 assumed certain quantities to be constant, and has given a more complete analysis free from these assumptions, using his analysis to solve the difficult

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problem of grid leak detection. His work represents an advance in theory, but the solutions of the circuit problems are presented in such a complex form that their application to design is not clear.

In general, the theoretical analysis has been confirmed experimentally at frequencies not greater than 500 kilo-cycles per second and only then for certain special cases. The assumptions made in most of the theoretical work regarding the constancy of certain quantities have not been subjected to experimental confirmation, and it is inadmissible to extend the existing theory to cases where the frequency is very high. No complete theoretical analysis of the single-stage amplifier with all types of anode load, taking account of all the factors involved, has been undertaken, and although the particular solutions rest on a sufficiently sound basis to enable the design of low and medium radio-frequency amplifiers to be undertaken with reasonable satisfaction, the present theory cannot be satisfactorily applied at very high frequencies. It is possible that small effects, such as the power factors of the inter-electrode capacities (which have hitherto been neglected, but which, as HARTSHORN has recently shown (see Abs. 1.2.7/9), may be appreciable), may play a part in the behaviour of the amplifier at high radio frequencies.

The design of short wave amplifiers is almost entirely empirical and, as it is known that the behaviour of a triode at very high frequencies presents features which have not yet been explained, theoretical analysis of radio frequency amplification cannot be considered as nearly complete. The growing utility of the short wave receiver demands a more rigid theoretical study of the subject.

1.2.3. MULTI-STAGE AMPLIFIERS

The previous remarks refer only to the analysis of the simple one-stage amplifier. When we consider the literature of multi-stage amplifiers, we find less definition.

A general examination of cascade amplifiers in which the effects of the anode-grid inter-electrode capacity are considered was made in an elementary manner by BARKHAUSEN (see Abs. 1.2.3/1) and in a more detailed manner by FORTESCUE (see Abs. 1.2.3/2). This latter work represents the collected results of experiment and constitutes an important step in the analysis of multi-stage amplifiers.

A careful theoretical analysis was made by OLLENDORFF in 1924 (see Abs. 1.2.3/6), who solved the circuit problem in terms of differential equations and thus made some progress towards the solution of the difficult problem of analysis under non-periodic conditions. The results are chiefly applicable to audio-frequency amplifiers but can be used for simple transient cases. OLLENDORFF'S theory was extended by H. KAFKA in 1925 (see Abs. 1.2.3/7), who also gives later an examination of inductive couplings between stages.

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1.2.4. Super-Regenerative and Supersonic-Heterodyne Receivers

The difficulty in analysing the properties of a super-regenerative receiver is that so many rather indeterminate factors have to be assumed. The properties outlined by Armstrong (see Abs. 1.2.4/1) are obtained by experiment only and the analysis given by Hulbert (see Abs. 1.2.4/7) is of an elementary nature only. A more complete discussion of such circuits is given by Appleton and Thompson (see Abs. 1.2.4/8), but in general it can be stated that very little exact theoretical information is available, in fact, David, Dufour and Mesny state (see Abs. 1.2.4/9 and /21) that such analysis cannot be rigorous at present, due to the lack of exact information regarding the constants involved. They adopt an experimental method, viz., an oscillograph to study the various types of oscillation which can be produced.

The problem of analysing the supersonic-heterodyne receiver is simpler, since it involves an analysis of the frequency transforming stage and the selectivity of the intermediate-frequency amplifier. The latter is dealt with under the general amplifier sections, while the former is more completely dealt with in Section II.

1.2.5. Radio-Frequency Filter Circuits

The theory and practice of Radio Filter circuits is so inextricably bound together that all references to this subject are contained in sub-section 1.1.9, to which the reader is referred.

1.2.6. Valve Noises and the "Schrot" Effect

W. Schottky (see Abs. 1.2.6/1) has predicted that since a thermionic electron current consists of the independent evaporation of successive electrons, the anode will be successively bombarded, giving rise to a rapid succession of impulses in the tuned anode circuit and thus giving a small output current by shock excitation when no input potential difference is applied. Experimental results have shown that this effect, known as the " Schrot " effect (1.2.6), definitely fixes an upper limit to the possible amplification of any valve system and is responsible for some of the inherent noise in multi-stage amplifiers. This output noise is due chiefly to the shock excitation of the anode of the first valve in a multi-stage system, since the output voltage due to this is amplified most. The total R.M.S. output depends upon the electron rate, i.e., the anode current, on the fundamental charge of the electron, on the capacity and power factor of the circuit and on the intensity of the space charge.

For ordinary radio receiving circuits, this voltage is of the order of 10–100 microvolts and sets an upper limit to the total useful amplification.

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The space current in a valve is not the steady flow of a continuous fluid. It is the migration of discrete particles which possess regularity only in the statistical sense. Either the discrete nature of the particles or the inequality in the time intervals between them would be sufficient to affect any receiving circuit that could distinguish between steady and variable currents. The magnitude of the effect might conceivably be very small, but if it were sufficiently amplified by a perfect amplifier it could be made to actuate a telephone receiver. It would then manifest itself as an audible noise. The intelligibility of a message received would depend upon the ratio of the signal to this effect. With perfect amplifiers, there is a lower limit of the field strength below which electrical signals cannot be detected by valves.

The " Schrot " effect will be measured by the difference between the energy actually dissipated in the measuring circuit, and the energy that could be dissipated if the electron stream were a uniform flow of a continuous fluid.

Published results agree fairly well with theory but do not extend to a lower frequency range, as shown by JOHNSON (see Abs. 1.2.6/3). At such frequencies, the experimentally determined value of the electronic charge is too large and suggests the need of some extension of the theory. JOHNSON suggests that this experimental difference is due to secular changes in the surface conditions of the cathode, and that the " Schrot " effect does not completely explain valve noises.

The effect is very serious in screen-grid valves of the Hull type, because of the excellent screening and enormous amplification.

A further source of amplifier noise has been examined by JOHNSON (see Abs. 1.2.6/10), which is due to the value of the resistance connected across the grid and filament of the first stage of an amplifier. The electric charges in a conductor are found to be in a state of thermal agitation, in therodynamic equilibrium with the heat motion of the atoms of the conductor. The manifestation of the phenomenon is a fluctuation of potential difference between the terminals of the conductor which can be measured by suitable instruments. The effect is one of the causes of that disturbance which is called " valve noise " in amplifiers. Indeed, it is often by far the larger part of the " noise " of a good amplifier. When such an amplifier terminates in a telephone receiver and has a high resistance connected between the grid and filament of the first valve on the input side, the effect is perceived as a steady rustling noise in the receiver, like that produced by the " Schrot " effect. The results of JOHNSON's work show that a low input resistance should be used if this " noise " is to be reduced and that the temperature of this resistance should be low.

An alternating potential of 10^{-6} volts marks a critical region for amplifiers, below which the valve " noises " will predominate.

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1.2.7. Input Impedance

Two papers of almost identical date—namely, 1920—due to Miller and Ballantine (see Abs. 1.2.7/2 and /4), have shown the importance of the input impedance of an amplification stage and the manner in which this is dependent upon the nature of the output anode circuit load and the coupling of this load to the input circuit by means of the anode-grid capacity. The valve network is analysed in both papers, and expressions for the resultant input impedance are given in the following form by Miller :—

(*a*) The input impedance depends upon the nature of the anode circuit load, owing to inter-electrode capacities.

(*b*) If the anode circuit load is a resistance or a capacity reactance, the input impedance is positive and absorbs power.

(*c*) When the load is inductive, the input impedance contains a negative resistance component, in which case the input circuit resistance is reduced. In the limiting case, when the power fed back to an oscillatory input circuit equals the losses in that circuit, self-oscillation occurs.

This statement of Miller is too general. The input resistance component may be positive with an inductive anode circuit load, as is shown by Miller's own curves ; this occurs when the anode circuit reactance is small. Reasonable experimental confirmation has been obtained at frequencies below 400 kilo-cycles per sec.

These important results have been applied to the more detailed study of amplification and to multi-stage amplifiers. The effective shunt loads that can be thrown across the anode circuit load of a preceding stage due to high-factor valves are in many cases very large.

The nature of the output load affects the amplification, and in fact the effect of the load is successively passed on to each stage of a cascade arrangement. No analysis of an amplifier that disregards the inter-electrode capacity coupling can be considered as even approximate at frequencies of the order of 1,000 kilo-cycles per sec.

Hartshorn (see Abs. 1.2.7/9) has recently shown that the valve capacities are not pure, but are associated with resistance components, i.e., the power factor of the inter-electrode capacities is not zero. These valve losses, due probably to hysteresis effects in the glass, will naturally modify the performance of the valve. A more detailed examination of the valve network has been made, and the effect of these conductances is only now being realised. It is to be noted that the measurements giving the value of the power factor were taken on an audio-frequency bridge. It is certain that the conductance value obtained at this audio-frequency cannot be applied to very high frequencies. In fact, there is evidence to support the idea that considerably greater losses than those shown by the bridge

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occur in the valve itself at frequencies above 10,000 kilo-cycles per sec.

In general, it can be stated that the close examination of the input impedance of the valve and its network has yielded far greater refinement in design, and that satisfactory agreement between theory and measurement exists at low radio-frequencies.

At very high radio-frequencies very little quantitative work has been done; moreover, there is a need for a much fuller investigation of the behaviour of the valve itself at frequencies above 10,000 kilo-cycles per sec.

1.2.8. Amplifiers Using Special Values

The utility of the "space-charge-grid" valve is to perform the same kind of functions as the triode, but at lower anode voltages or with somewhat higher amplification. As an example, the following details of such a valve are given (see Abs. 1.2.8/14) when used as an amplifier :—

Coupling resistance	300,000 ohms.
Grid Bias	—0·75 volt.
H.T. Supply	180 volts.
Actual Anode Voltage	90
Amplification Factor	80
Internal resistance of Valve	160,000 ohms.

A triode having similar electrodes and an amplification factor of 80 and worked at an anode voltage of 90 would have a mutual conductance of about one-fourth as much as the space-charge-grid valve.

One of the limitations in the use of valves with a very high amplification factor in resistance-capacity-coupled circuits is that their effective input capacity is apt to be very high. This results in a falling off of amplification at the higher audio frequencies. The higher the amplification of the stage, the more serious this effect becomes. This limitation applies to the space-charge-grid valve having a high amplification factor. Valves with a factor of 6 to 10 have been used quite extensively in Europe for a number of years. European practice has been to design and use the valves in such a way as to permit very low anode voltages of the order of 6 to 20 volts.

The growth of the screen-grid tetrode has given rise to several discussions of its properties, notably those of Hull (see Abs. 1.2.8/3), Beatty (see Abs. 1.2.8/6) and Warner (see Abs. 1.2.8/14).

The properties of a perfect valve of this type are that the anode grid capacity is zero and that the internal resistance of the valve is infinite. In practice, it is impossible to obtain these values, the nearest approach being that due to Hull, in which the anode-grid capacity was 0·006 $\mu\mu$ F. and the internal resistance was

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5×10^5 ohms. The voltage amplification falls as the frequency is increased, being 7 at 10,000 kilo-cycles per sec.

In commercial valves in this country such ideal conditions are not obtained, an anode-grid capacity of about $0 \cdot 1$ $\mu\mu$ F. being usual with an internal resistance of $1 \cdot 25 \times 10^5$ ohms, the voltage amplification being about 13 at a frequency of 4,000 kilo-cycles per sec. The general properties of the present commercial types are given by McLACHLAN (see Abs. 1.2.8/8) and HALL (see Abs. 1.2.8/10).

It should be noted that the primary object of the screen-grid tetrode is to reduce the anode-grid capacity, thereby eliminating retro-action difficulties, but it has recently been shown that even with such small inter-electrode capacities as $0 \cdot 1$ $\mu\mu$ F. " neutralisation " should be used if the full voltage factor is to be utilised. The high voltage factor is incidental to the design and is not the original object of the development of this valve.

1.2.9 SCREENING OF RADIO-FREQUENCY AMPLIFIERS

Two distinct types of screen may be employed to shield radio apparatus :—

(*a*) A screen to prevent electric or magnetic induction from one part of the receiver to any other part.
(*b*) A screen to prevent direct external pick-up.

The former type of screen is used to isolate coils or stages and the latter to enclose the complete receiver. Owing to the weak fields produced by a distant station, this latter type of screen does not need to be so perfect as the former type.

In both cases we have two types of field to consider, viz., the magnetic and the electric. In the magnetic case a metal cage is used surrounding that portion of the receiver to be screened. In this case the efficiency of the screen depends upon the eddy currents produced in the shield. When the resistivity and thickness of the metal remain constant and the frequency of the alternating electromagnetic field varies, the shielding increases as the frequency increases, and when the frequency and resistivity remain constant, the shielding increases as the thickness is increased.

The penetration of the eddy currents into the conductor will be governed by the ordinary well-known " skin-effect " formula, and hence the latter will determine the fraction of the original magnetic field which penetrates to any depth in the conductor. The decay factor governing the amplitude of the radio frequency current at a depth " x " below the surface of the metal is $e^{-2\pi\sqrt{\mu f/\rho}\cdot x}$ where " μ " and " ρ " are the permeability and resistivity of the metal respectively and " f " is the frequency of the alternations (see Abs. 1.2.9/1). Using this formula, the reduction in the field strength can be calculated for a screen of any given thickness. It can easily

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be shown that iron is 5·5 times more effective than copper if its thickness is the same. It is also clear that the screening effect is greatest at very high frequencies and that at audio frequencies effective screening becomes increasingly difficult.

Magnetic shielding is always accompanied by an increase in the effective resistance of the shielded circuit ; for a given kind of metal at any specified frequency, a certain thickness of shield introduces a maximum resistance into the shielded circuit. The thickness of shield which gives maximum added resistance to the shielded circuit decreases as the frequency increases. If shielding is to be obtained by eddy currents, they must be free to flow as they will. Any imperfect joint in the shield or slits, which tend to constrain the eddy currents to restricted paths, will seriously interfere with the shielding obtainable. The smallest holes or cracks in a screen will permit a detectable amount of the field to escape.

A short circuited coil may be used as a shield, since the current induced in it by the field will set up an opposing field and give a zero local resultant. In this case, the shielding is zero when either the inductance of the shielding coil or the frequency of the alternating electromagnetic field is zero and is perfect when the resistance of the shielding coil is zero.

In the case of shielding from electric fields, any space completely surrounded by conducting material is completely shielded if the field is constant, but if it is changing, the space inside the cage is not completely shielded, since the induced charges are moving and do not take up their steady-state positions immediately. Thus a low resistance screen is required to give the best screening from alternating electric fields. If no energy is to be absorbed in the screen, the conductors in the screen must be " open " circuits, e.g., insulated wires or open loops, so that no eddy currents can be produced in them. In this latter case, the electric shield can be used to prevent the alternating field from reaching an impure dielectric and thus producing a loss. In this capacity the shield reduces the effective resistance of the electrical circuit.

Although it is practically impossible to obtain perfect screening, it is comparatively easy to obtain sufficient screening for most practical purposes. Such a coarse mesh screen as can be obtained with wire netting with spacings of 1 inch between conductors can give 95 per cent. efficiency at 1,000 kilo-cycles per sec., and such screens are usually adopted for large volumes. For receiving apparatus, metal boxes are advisable, and care should be taken to reduce the number of holes to a minimum.

1.2.10 Interference at Radio Frequencies

There are many types of interference to which a radio-frequency system may be subjected. The most important are :—

(*a*) Interference from other transmitters working on neighbouring frequencies.

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(*b*) Interference from atmospherics.
(*c*) Local interference due to induction from disturbing influences such as railways, transmission lines, electric supply mains, etc.

Treating these three main causes independently, we will first consider case (*a*).

Interference from other transmitters is due to the selectivity characteristic of the receiver being such that the interfering station can produce a weak current in the tuning system of the receiver. Selectivity is that characteristic of a receiver which enables it to discriminate between two incoming signals on neighbouring frequencies, passing one and excluding the other. It involves a progressively increasing attenuation of the radio frequency tuning system towards frequencies increasingly removed from the desired signal frequency. It should be noted that in the case of telephony reception the total width of the side bands must be considered in stating the frequency spectrum occupied by a particular station. The obvious method of eliminating this source of interference is to sharpen the response characteristic of the receiver, this usually being done by the use of retro-action. However, if the selectivity characteristic is too peaked, a side band cut-off with its consequent distortion will be produced. The supersonic-heterodyne receiver is far more selective than a normal two-tuned-circuit neutrodyned receiver, and is better for this purpose, although its selectivity characteristic can be made so good that serious side-band cut *off* is obtained. A band-pass filter should be used in this case, since by its use a flat-topped response characteristic can be obtained. Another method is to use several slightly dissimilar tuned circuits for the intermediate stages of the supersonic-heterodyne receiver, the combination of which gives a flat-topped overall characteristic.

A powerful interfering signal will also affect the sensitivity of a receiver tuned to a weak neighbouring signal (see Abs. 1.2.10/6). In general such interference will reduce the sensitivity due to a reduction in the normal retro-action. The interference produced by spark stations is far greater than that produced by C.W. stations.

The interference produced by atmospherics (*b*) forms one of the most important practical limitations to high amplification in practice. The essential factor which sets an upper limit to useful amplification is the ratio of the signal intensity to that of the atmospheric disturbances.

Owing to the transient nature of atmospheric disturbances, no receiving system, however selective, can be made unresponsive to such disturbances, and BELLESCIZE (see Abs. 1.2.10/11) concludes that at small or moderate distances the atmospheric produces a disturbance which is equally effective whatever the wave-length to which the receiver is tuned. Hence an increase of the amplification or of the sensitivity of the receiver would simply lead to an equal

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relative increase of both the signal and the atmospheric with no effective gain produced by amplification.

Various attempts to eliminate atmospheric disturbance effects have been made (see Abs. 1.2.10/4), but it seems generally adopted that the only methods which can usefully be employed are those which involve the directional discrimination of the signal and atmospheric and those which employ band filters between the antenna system and the receiver (see Abs. 1.2.10/3). Many successful forms of directive systems are in use which can materially reduce the effect of atmospherics by so adjusting the system that practically no pick-up exists in the direction of the atmospherics. Unfortunately, however, such systems break down if the direction of the atmospheric disturbance is not precise and covers any considerable angle.

It has been shown (see Abs. 1.2.10/5) that the signal to interference ratio in single side-band transmission without the carrier wave is double that obtained by the use of double side-band transmission, owing to the narrower frequency band over which the energy is distributed.

The last type of interference (*c*) is not very prominent in most cases. The effects of local interfering magnetic and electric fields produced by lighting mains and electrical apparatus in general are usually confined to the audio-frequency section of the amplifier where they may be serious. However, high harmonics of lighting supply frequencies and radio-frequency interference from damped oscillations set up by local arcing and sparking are not uncommon (see Abs. 1.2.10/12). Such interference can usually be reduced by the judicious use of chokes and wave traps or rejector circuits (see Abs, 1.2.10/8).

BIBLIOGRAPHY

1.2.1. General Theory and Transients

1. **Electrical Conductivity produced by Hot Conductors :** O. W. Richardson : *Phil. Trans. Roy. Soc.*, 1903, **201**, 497–549.

Platinum, carbon, or sodium is heated to redness in a vacuum, and the negative leak from it is studied. On the corpuscular theory of conduction in metals a formula of the type $A\,\theta^{\frac{1}{2}}e^{-\frac{b}{\theta}}$ is deduced as giving the current at a temperature θ, and this is found to fit the experimental results. From the constants A and b involved, the number of corpuscules per c.c. can be deduced, as well as the work required to drag a corpuscule from the metal. Corpuscular currents of the order of 1 amp. per sq. cm. of area were obtained in the case of carbon. The experiments completely explain the Edison effect in incandescent lamps. At high temperatures, the conductivity of the vacuum may be quite comparable to that of the filament.

This is Richardson's original classical paper on Thermionics.

2. **Effect of Space Charge and Residual Gases on Thermionic Currents in High Vacuum :** I. Langmuir : *Phys. Rev.*, 1913, **2**, 450-86.

In this classical paper the fundamental laws of electron emission in the diode are given, and experimental results show that Richardson's equation

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is valid at the lower temperatures but at the higher temperatures the current becomes independent of the temperature and dependent upon the anode voltage. LANGMUIR then enunciates his 3/2 power law and also determines more accurately the values of "*a*" and "*b*" in Richardson's equation. He concludes with the assertion that the emission of electrons from an incandescent solid in a vacuum below 10^{-6} mm. is a specific property of the substance and is not dependent on the gas, as had hitherto been assumed.

3. **The Pure Electron Discharge and its Application in Radiotelegraphy and Telephony :** I. LANGMUIR : *Gen. Elect. Rev.*, 1915, **18**, 327-39 ; *Electrician*, 1915, **75**, 240-45 ; *Elect. Rev. Lond.*, 1915, **76**, 753-55 ; *Lumière Élect.*, 1915, **29**, 241-45, 272-79.

This important paper shows the development of the triode, the general effect of the third electrode upon the current in the anode circuit being discussed. The necessity of removing occluded gases from the electrodes and the inner glass surfaces of the bulb is pointed out and the difficulty of construction is shown not to be so much in the production of a high vacuum but rather in the maintenance of this vacuum during the life of the valve. As the voltage applied to the anode is increased, the tendency for the gas residue to become ionised is more marked. The most sensitive indication of the presence of ionisation is the failure of the current to increase with the voltage in a regular manner.

4. **General Expansion Theorem for the Transient Oscillation of a Connected System :** J. R. CARSON : *Phys. Rev.*, 1917, **10**, 217-25.

In the usual solution of the problem of the transient oscillation of a connected mechanical or electrical system in response to a suddenly impressed set of forces, the determination of the characteristic modes of oscillation (periodicities and dampings) is comparatively easy, since it involves only the determination of the roots of a polynomial. But the determination of the amplitudes of the transient oscillations is extremely laborious, and the difficulty increases rapidly with the number of degrees of freedom of the system. HEAVISIDE, in his "expansion theorem," has given a valuable formulation of the transient oscillation of an electrical network when the oscillations are excited by the sudden application to the system of an E.M.F. which is not a function of time, i.e., a steady uniform E.M.F.

In the present paper, the general solution of the problem considered by Heaviside is developed, and an expansion theorem is derived which formulates the resultant (forced and transient) oscillations of a connected dynamical system in response to an arbitrary driving force applied to any co-ordinate of the system. The only limitation imposed on the form of the driving force is that it shall be capable of expansion in a Fourier's integral or else be expressible as a complex exponential function of time, a limitation which constitutes no restriction of practical importance. The chief utility of the theorem rests in the fact that by its use the solution for the transient oscillation of the system considered is reduced to formulæ which are functionally the same as those for steady-state oscillations.

5. **Generalisation of Heaviside's Expansion Theorem :** A. PRESS and J. B. POMEY : *Rev. Gen. Elect.*, 1918, **4**, 691-94.

6. **Theory of the Transient Oscillations of Electrical Networks and Transmission Systems :** J. R. CARSON : *J. Amer. Inst. Elect. Eng.*, 1919, **38**, 405-89.

A simple formula is derived which expresses the transient current in terms of two independent functions : one, the applied E.M.F. expressed as a time function, and the other, a characteristic function of the constants and connections of the system, this latter being termed the "indicial

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admittance " of the system. A knowledge of the indicial admittance, therefore, completely determines the behaviour of the system to all types of applied voltages including both transient and steady states. A systematic investigation of methods for solving and computing the indicial admittance follows, in the course of which original solutions for transmission and artificial lines are derived and a new method involving integral equations is developed. This latter is particularly well adapted to handle the difficult problem of taking into account the effects of terminal impedances in transmission systems.

The paper is very comprehensive and gives accurate mathematical solutions of the problems dealt with.

7. **Amplification Constant and Internal Anode-Circuit Resistance of a Three-Electrode Valve :** J. M. MILLER : *Proc. Inst. Radio Eng.*, 1920, **8**, 64-73.

In this paper, formulæ based on theoretical considerations are derived, showing the dependence of the amplification constant and the internal anode circuit resistance of a valve on the dimensions and spacing of the three electrodes. A set of curves based on the derived formulæ allow of the rapid pre-determination of the amplification constant when the physical dimensions of the valve are known.

(See also Abs. 4.1.1/2.)

8. **Notes on the Theory of Modulation :** J. R. CARSON : *Proc. Inst. Radio Eng.*, 1922, **10**, 57-64.

The transmission system of " frequency modulation " (transmission by variation of the frequency of the radiated wave) is analysed mathematically and the width of the frequency band occupied by this method of transmission is compared with the width of the corresponding band for transmission by amplitude variation. It is proved that the frequency modulation system is inferior to the amplitude variation system both as to the width of the frequency band and as to the distortion of the signal wave form.

9. **Application to Radio of Wire Transmission Engineering :** L. ESPENSCHIED : *Proc. Inst. Radio Eng.*, 1922, **10**, 344-68.

The transmission losses in wire and radio communication are compared for various frequencies and distances of transmission. The effect of natural and signal interference at the receiving ends of such systems is described, and conclusions are drawn as to the necessary minimum transmission levels to be maintained in wire and radio circuits. The limitations on two-way operation resulting from oscillation of the entire system are considered in each case, and for combination wire and radio circuits as well. The comparative usefulness of repeater stations for wire and radio telephony is described. Several means of effective wire and radio telephony are described, notably, transmission on one side-band only. The transmission requirements of a radio telephone system are then given in numerical form.

10. **Heaviside's Expansion Theorem :** L. COHEN : *J. Franklin Inst.*, 1922, **194**, 765-70.

The theorem gives the distribution of current in any network, however complex, for both the steady and the transient states. The greatest advantage of employing the theorem is, of course, in the determination of transient effects, which, by the usual methods, is at best a laborious process, the difficulties increasing as the number of degrees of freedom of the system increases. The author has worked out the derivation in detail for a constant E.M.F. = E, and a sinusoidal E.M.F. = $E\varepsilon^{j\omega t}$ (frequency = $\omega/2\pi$), in a simple and easily understood form, while retaining Heaviside's expansion methods and, incidentally, explaining them.

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11. **On a Type of Oscillation-Hysteresis in a Simple Triode Generator :** E. V. APPLETON and B. VAN DER POL : *Phil. Mag.*, 1922, **43**, 177-93.

Deals with the " Ziehen " effect of German authors, to which the present authors have given the more applicable term " oscillation hysteresis." A general theory is developed to determine the possible and stable amplitudes when the characteristics are given. The conditions necessary for two stable amplitudes are examined, and it is found that for two stable amplitudes, one of which is zero, the expression for the characteristic when developed as a Maclaurin's series must have positive fifth and negative third differential co-efficients. Experiments confirming this are described, and it is shown that the relation between the oscillation amplitude and one of the circuit constants is irreversible, so that, for certain values, the amplitude may be either zero or finite according to the method of approach. It is shown theoretically that an electrical impulse above a limiting value may cause the amplitude to jump from zero to a finite value. This is proved experimentally by the use of a transient E.M.F. from a coil and moving bar-magnet. The relations of the critical values for the starting and stopping of oscillations and the maintained electrode potentials are investigated experimentally, and it is found that there is a region of grid potential in which the amplitude is a single-valued function of the circuit constants, so that conditions for starting and stopping are the same. For such values of grid potential, the characteristic is approximately represented by a power series of three terms. For other values of grid potential a difference in critical circuit constants is found, with possibility of the hysteresis effect. A power series of five terms for the characteristic accounts for the phenomena. Experiment shows that two stable amplitudes, both differing from zero, are possible, depending upon the seventh differential of the characteristic.

12. **Spectrum Analysis of Radio Signals :** C. SNOW : *Sci. Pap. Bur. Stand. Wash.*, 1923, No. 477, 231-61.

It is shown that the distribution of energy over a band of frequencies can be obtained by measuring the current in a tuned circuit which is successively tuned to a series of frequencies covering the energy band. This, however, is only true if the examining circuit has a relatively high resolving power, compared to the spectral structure of the wave. When this is not true, the resonance curve of the examining circuit must be considered in conjunction with the form of the energy band under investigation. Four mathematical methods of solving this problem are given.

13. **Relations of Carrier and Side-Bands in Radio Transmission :** R. V. L. HARTLEY : *Proc. Inst. Radio Eng.*, 1923, **11**, 34-56.

The modulation of a carrier wave by a signal wave and the subsequent reproduction of the signal wave are discussed in terms of the sustained sinusoidal components into which the various waves may be resolved. The conception of a modulated wave as being made up of a carrier component and two side-bands is explained.

The saving in frequency range resulting from transmitting a single side-band is emphasised. Carrier suppression and homodyne reception are shown to save power at the transmitter, increase the sensitivity of the receiver and aid in suppressing interference from both stations and atmospherics.

Equations are derived for the distortion of the reproduced wave in terms of that of the modulated wave. In general, both the amplitude and phase of each component of the signal depend on both the amplitudes and phases of the corresponding side-band components and of the carrier wave.

The effect on the signal of various typical distortions of the radio wave is examined for both single and double side-band transmission, as is also

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that of altering the phase of the locally supplied carrier and of altering its frequency. The resulting distortion of the signal is found, in general, to be more serious for telephony when both side-bands are used and for telegraphy when only one is used.

It is also shown that the reproduced signal has twice the power when both side-bands are transmitted as when the same transmitted power is concentrated in one. The signal to atmospheric ratio is, however, the same in both cases.

14. **Transient Oscillation in Electric Wave-Filters :** J. R. CARSON and O. J. ZOBEL : *Bell System Tech. J.*, 1923, **2,** 1-52.

(See Abs. 1.1.9/4.)

15. **Applications of Heaviside's Expansion Theorem :** L. COHEN: *J. Franklin Inst.*, 1923, **195,** 319-26.

The author points out that in some special cases it may be simpler to transform the Heaviside formula (see Abs. 1.1.9/4) into a form derived from other methods, while in others the only practicable method is to apply the formula directly. Both methods are illustrated in the four problems dealt with.

16. **Valve Amplifiers :** F. OLLENDORFF : *Arch. Elektrotech.*, 1924, **13,** 274-91.

(See Abs. 1.2.3/6.)

17. **Some Thermionic Valve Problems :** E. V. APPLETON : *Rep. Brit. Ass.*, 1925 ; *Electrician*, 1925, **95,** 266-67.

The author suggests that, to explain the shape of the static characteristics of thermionic valves, it must be assumed that all the electrodes emit electrons, and that, under certain conditions, the anode emission, due to electronic bombardment, may be almost as high as the filament emission. The amplification factor is the ratio of the number of lines of electric force ending on the grid to the number ending on the anode when grid and anode are at the same potential. An experimental bridge method of measuring amplification is outlined. Regarding the lag in the action of thermionic valves, it is found that as the time of applied voltage is increased, the internal resistance increases, reaching a constant value for a steady potential ; this is due to the alteration of the thermal condition of the filament, consequent on the closing of the anode circuit. The limiting conditions for the starting and stopping of oscillations are often not the same ; this type of oscillation hysteresis is due to the shape of the valve characteristics. The possibility of making a receiver oscillate with one of the possible and stable amplitudes for the same operating conditions constitutes a second type of oscillation hysteresis and is due to damping in the grid circuit.

18. **On the Theory of the Triode Valve Amplifier and the Measurement of the Characteristic Constants of the Triode Valve :** H. NUKIYAMA and S. KITTA : *J. Inst. Elect. Eng. Jap.*, 1925, No. 445, 664-78.

In this paper the theory of the dynamic characteristics of a triode valve is based on the assumption that the alternating currents in the grid and the anode circuits can be expressed as complex linear functions of the alternating potential applied to them. A general theory to determine the four valve constants is given. In a particular case, the valve constants are reduced to three, i.e., input impedance, anode circuit resistance and voltage factor.

(See also Abs. 4.1.1/4.)

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19. **The Inter-Electrode Capacities of Thermionic Valves :** L. HARTSHORN and T. I. JONES : *Experimental Wireless*, 1925, **2**, 263-73.

The upper limit of the frequency of a valve oscillation generator is governed by the valve capacity, and this limit is calculated for various types of oscillation circuit.

It is shown that the inter-electrode capacities are always impure, being associated with a conductance. Power factor measurements have been taken to ascertain the value of these conductances. Finally, the effects of various types of valve base are discussed.

(See also Abs. 4.1.3/4.)

20. **" Relaxation " Oscillations :** B. VAN DER POL : *Jahrb. drahtl. Telegr.*, 1926, **28**, 178-84, and *Phil. Mag.*, 1926, **2**, 978-92.

In the first part of this paper the author develops the theory relative to oscillations in systems with a single degree of freedom and including a negative resistance. Conditions are obtained for the solution of the appropriate differential equation to be aperiodic, periodic, or quasi-aperiodic. For the last case, the solution obtained differs considerably from the normal approximately sine-formed solution, a pure periodic solution being found whose period is expressed by a " relaxation time " (time constant) of the system. On this account the term " relaxation oscillations " may be given to such phenomena. It is shown that the oscillations of multivibrators of the Abraham and Bloch type are examples of a general type of relaxation oscillations. It is also probable, though not yet strictly proved, that the oscillations of a Wehnelt interrupter belong to the general class of relaxation oscillations.

21. **Four-Element Valve Characteristics :** D. C. PRINCE : *Proc. Inst. Radio Eng.*, 1928, **16**, 805-21.

The variation in the ratio of the grid to the anode current of three-electrode vacuum valves aroused considerable curiosity, and tests were undertaken in an endeavour to ascertain the laws of current division. It was found that in a valve having symmetrical electrodes—that is, straight filament, concentric cylindrical anode and cylindrical grid, made of wires parallel to the axis—the ratio of the grid to the anode current was a function of the valve geometry and quite different from that usually found in commercial designs. The ratio of the grid to the anode current in such a valve exceeds the ratio of the projected grid area by only a small amount, easily accounted for by variations in the electric field around the grid wires. The considerable departures from this ratio in commercial valves appear to be due to a combination of secondary emission from the valve anode and unsymmetrical arrangements of the grid wires and supports.

22. **Unstable Equilibrium and Parasitic Phenomena in Triode Circuits :** M. PODLIARSKY : *Onde Élect.*, 1928, **7**, 287-306.

The emission of secondary electrons in a triode may lead to an inversion of anode or grid current. When a valve showing inversion is associated with a high resistance, the characteristics of the whole circuit may give rise to two possible states—normal and parasitic—and the behaviour may be uncertain. The author examines the conditions which result in this instability both in the case of diodes and triodes, and suggests precautions that may be taken to prevent parasitic states, which are frequently destructive.

23. **Practical Application of the Fourier Integral :** G. A. CAMPBELL : *Bell System Techn. J.*, 1928, **7**, 639-707.

The growing practical importance of transients and other non-periodic phenomena makes it desirable to simplify the application of the Fourier

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integral in particular problems of this kind and to extend the range of problems which can be solved in closed form by this method. To facilitate the use of the known closed-form evolutions of Fourier integrals, many of them have been compiled and are tabulated. They are presented, however not as definite integrals, but as paired functions, one function being the co-efficient for the cisoidal oscillation (or complex exponential) and the other function the reciprocally related co-efficient for the unit impulse. This arrangement simplifies the table and promises to be most convenient in practical applications, since it is the co-efficients of which immediate use is made, just as in the case of the Fourier series. Applications of the tabulated coefficient pairs to 85 transient problems are given, together with all necessary details.

24. **On the Distortionless Reception of a Modulated Wave and its Relation to Selectivity :** F. K. VREELAND : *Proc. Inst. Radio Eng.*, 1928, **16**, 255-80, Disc. 494-97, 671-73, 848-50.

The author first enters into a discussion of the various factors which affect the final reproduced acoustic output from a radio receiver. He shows that if the hollow tone associated with a preponderance of low audio-frequencies is to be avoided, the overtones so necessary for fidelity must be preserved. These overtones are transmitted by the side-bands, and if we assume that the modulation includes frequencies up to 10,000 cycles per second, the total width of the transmitted spectrum, including both side-bands, will be 20 kilo-cycles per sec. The two side-bands are not both necessary for distortionless transmission. Either may be suppressed without affecting the fidelity of reproduction, and in some transmissions one side-band is completely suppressed, but when both are transmitted, each must be received in its entirety.

The condition of distortionless reception and faithful reproduction of tone is that all the frequencies in the transmitted band of the modulated wave shall be received in their true relative intensities. On the other hand, the condition of selectivity requires the effective separation of all the frequencies of an interfering wave from those of the desired signal ; obviously, if the side-bands of the two interfering waves overlap, the two conditions of fidelity and selectivity are incompatible.

The ideal solution of this problem is to use a system which has a substantially rectangular frequency characteristic, that is, one which gives substantially uniform reception over a definite band of frequencies, including all the side-band frequencies of the modulated wave, with a cut-off for frequencies outside this band.

The problem permits of two distinct solutions :—

(i) The use of one, or more, band selectors, each of which possesses a substantially rectangular frequency characteristic.
(ii) The use of two, or more, receiving elements whose individual frequency characteristics are not rectangular but which, in combination, have an overall characteristic which is substantially rectangular.

Examples of the former type are given, together with the oscillograms obtained by means of a cathode ray oscillograph. The latter solution can be given approximately by the use of two, or preferably three, stages, the resonant frequencies of which are slightly dissimilar. The addition of three identical but spaced resonance curves gives a sensibly rectangular frequency response characteristic. The author concludes by giving oscillograms obtained from amplifiers of this type, from which it can be seen that the rectangular response characteristic can be closely approximated to in practice.

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1.2.2. The Valve Network

1. **Characteristic Curves and the Conditions for Stable Operation :** J. Béthenod : *Lumiere Élect.*, 1916, **35**, 25-31.

The paper discusses the elementary conditions necessary for a valve network to remain stable.

(See also Abs. 1.4.1/1.)

2. **Discussion of the Audion :** M. Latour : *Lumiere Élect.*, 1916, **35**, 289-92.

A simple theory of the single valve as an amplifier is given in terms of the four parameters of the valve. The optimum value of anode resistances is calculated and the use of a tuned anode circuit is suggested. In this latter case, the conditions under which this tuned system acts as a pure ohmic resistance are expressed and the behaviour of a valve oscillating system with grid to anode coupling is examined.

3. **The Working of the Three-electrode Vacuum Tube in Wireless Telegraphy :** G. Vallauri : *R. Accad. Navale, Livorno, Instit. Elettrotecn. e. Radiotelegrafico. Publ. No.* 1. *Elettrotechnica*, 1917, **4**, 43-50, 63-70.

An analytical attempt is made to examine the properties of a valve as an amplifier, rectifier and oscillator. The two circuits, namely, grid-filament and anode-filament, are considered as coupled inside the valve, so that either of the currents in these circuits can be considered as functions of the two potential differences between grid and filament and between anode and filament.

(See also Abs. 2.3.2/1.)

4. **General Expansion Theorem for the Transient Oscillations of a Connected System :** J. R. Carson : *Phys. Rev.*, 1917, **10**, 217-25.

(See Abs. 1.2.1/4.)

5. **Theory of the Thermionic Amplifier :** H. J. van der Bijl : *Phys. Rev.*, 1918, **12**, 171-98.

This paper is an excellent early attempt to investigate the properties of an amplifier, the fundamental amplification equations being derived and the correct use of the static characteristics suggested.

The case of a pure resistance load only is considered and the bulk of the paper is devoted to a study of the valve itself rather than to an examination of the properties of an amplifier.

6. **Generalisation of the Heaviside's Expansion Theorem :** A. Press and J. B. Pomey : *Rev. Gen. Elect.*, 1918, **4**, 691-94.

7. **Vacuum Tubes and their Technical Applications :** H. Barkhausen : *Jahrb. drahtl. Telegr.*, 1919, **14**, 27-47 and **16**, 82-114.

The first part of this paper gives a full discussion of the action and design of three-electrode valves, and their use in the construction of amplifiers. The second part deals with the theory of amplifiers, including : (1) the effect of the input transformer used with the first valve ; (2) the effect of the capacity between grid and anode ; (3) amplification ratio ; (4) retro-action coupling ; (5) limits of amplification in multi-stage amplifiers.

The paper deals chiefly with the theoretical design of the valve, the analysis of its behaviour in a circuit being elementary.

(See also Abs. 1.4.2/3.)

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8. **A Theoretical Study of the Three-Element Vacuum Tube :** J. R. CARSON : *Proc. Inst. Radio Eng.*, 1919, **7**, 187-200.

Starting with a functional relation between the anode-current and the grid and anode voltages, the anode-current variation is deduced to a sufficient degree of approximation in terms of the constants of the valve and the grid-voltage variation, for pure resistance output loads only.

The requirements of amplification, modulation and detection are then deduced from the equations obtained. Heterodyne amplification is discussed and it is held that, within the working range of the valve, this amplification is the ratio of the amplitudes of the locally and remotely generated waves, provided the latter is small as compared with the former. The general problem of the output current variation with an impedance in the output circuit is then considered and the production of harmonic frequencies observed.

Although inter-electrode capacity effects are neglected, the voltage factor and internal resistance of the valve are assumed constant and the effects of grid current are neglected, this paper represents a definite stage of theoretical progress. It should be noted, however, that the premises are too elementary for any exact deductions.

(See also Abs. 2.3.2/6.)

9. **The Audion as a Circuit Element :** H. W. NICHOLS : *Phys. Rev.*, 1919, **13**, 404-14.

The functional equations of the triode are used to deduce the actions taking place in any circuit containing a triode. The equations for small variations in the general network containing triodes are developed and simple equivalent circuits to replace the valve are shown.

(See also Abs. 1.2.7/1.)

10. **Dependence of the Input Impedance of a Three-Electrode Vacuum Tube upon the Load in the Plate Circuit :** J. M. MILLER : *Sci. Pap. Bur. Stand. Wash.*, 1919, **15**, No. 351, 367–85.

In this important paper, the valve circuit network is carefully examined, formula of a practical form being given for the calculation of the input capacity and resistance, here considered to be in series. Satisfactory experimental confirmation is given. It is shown that a resistance or capacity anode circuit load produces a positive input resistance, and an inductive load a negative resistance.

(See also Abs. 4.1.3/1.)

11. **Design of Multi-Stage Amplifiers :** C. L. FORTESCUE : *J. Inst. Elect. Eng.*, 1920, **58**, 65-74 ; *Radio Rev.*, 1920, **1**, 178-81 : *Elect. Rev. Lond.*, 1920, **85**, 700-1.

(See Abs. 1.2.3/2.)

12. **The Unilateral Dynamic Characteristics of Three-Electrode Vacuum Tubes :** J. G. FRAYNE : *Phys. Rev.*, 1922, **19**, 629-51.

(See Abs. 3.4/1.)

13. **Heaviside's Expansion Theorem :** L. COHEN : *J. Franklin Inst.*, 1922, **194**, 765-70.

(See Abs. 1.2.1/10.)

14. **On a Type of Oscillation Hysteresis in a Simple Triode, Generator :** E. V. APPLETON and B. VAN DER POL : *Phil. Mag.*, 1922, **43**, 177-93.

(See Abs. 1.2.1/11.)

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15. **Applications of Heaviside's Expansion Theorem :** L. COHEN : *J. Franklin Institute,* 1923, **195,** 319-26.

(See Abs. 1.2.1/15.)

16. **High Frequency Amplifiers :** H. T. FRIIS and A. G. JENSEN : *Bell System Techn. J.,* 1924, **3,** 181-205.

(See Abs. 1.2.3/5, 4.2/6.)

17. **Operation of Thermionic Vacuum Tube Circuits :** F. B. LLEWELLYN : *Bell System Techn. J.,* 1926, **5,** 433-62.

LLEWELLYN states that the earlier work summarised very concisely in an important paper by CARSON (see Abs. 1.2.2/8) assumes certain quantities to be constant, and that there is a need for more detailed examination in which the effects of the variation of the quantities are considered. He has, therefore, given a more complete analysis of the valve network and the results of such a closer examination are applicable to the problem of rectification. Given the static characteristic of grid-current grid-potential and anode-current anode-potential for any triode, the general exact equations are derived for the output current when the valve is connected in circuits of any impedance whatsoever and excited by any variable voltage. The method of derivation is illustrated in the special case where resistances only are considered and the adaptation of complex impedance to use in non-linear equations is shown. Approximations that are allowable in various practical applications are indicated, and the equations applied in some detail to grid-leak rectifiers, and in brief to other types of rectifiers, modulators, amplifiers and oscillators.

The theoretical results obtained are presented in such a form as to give very little clear indication of design requirements and are best adapted to the solution of abnormal cases.

The paper represents a vast improvement upon any earlier work, but is rather too mathematical for general design purposes.

(See also Abs. 2.3.2/17.)

18. **The Amplification and Selectivity of a Neutralised Tuned Anode Circuit :** N. W. MCLACHLAN : *Experimental Wireless,* 1926, **3,** 545–52.

The paper gives a simple discussion of the amplification and selectivity of a neutralised tuned anode circuit, the conclusions arrived at being as follows :—

(1) The stage factor K cannot exceed μ, the valve voltage factor.
(2) K increases for a given λ as L increases.
(3) K decreases for a given λ as C increases.
(4) If μ is constant K decreases as r_p increases.
(5) K increases as μ increases if r_p is constant.
(6) Small L with low R and large C give selectivity.
(7) But R is often swamped by grid damping of next stage.

These conclusions are obtained by assuming perfect neutralisation. It is doubtful whether such theoretical conditions can be obtained in practice and, therefore, the analysis of a neutralised stage is more complex than this paper would indicate.

19. **Determination of the Optimum " Durchgriff " $\left(\frac{1}{\mu}\right)$ for Amplifying Valves :** A. FORSTMANN and E. SCHRAMM : *Jahrb. drahtl. Telegr.,* 1927, **30,** 89-95.

The analysis given shows the working equations for a valve when the anode load is either a pure resistance or an impedance. A method is

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thus deduced which, allowing for the limitations of rectilinearity of characteristic and of internal resistance, gives the optimum value for the "Durchgriff" $\left(\frac{1}{\mu}\right)$.

The analysis assumes the valve characteristics to be linear and neglects the effects of inter-electrode capacities.

20. **Tuned Radio Frequency Amplifiers :** R. S. GLASGOW : *J. Am. Inst. Elect. Eng.*, 1928, **47**, 327-31.

A simplified method of calculating the performance of a tuned amplifier circuit is given and the effect of the circuit constants on the amplifier characteristics is shown. The effect of the coupling of the radio frequency transformer upon the selectivity is analysed and a few experimental results are given. These results are not in close agreement with the theoretical curves and the analysis naturally disregards minor circuit details. The paper concludes with a discussion of various types of neutralising devices to render a receiver stable over a large range of frequencies.

21. **The Calculation of Medium Frequency Amplifiers for Super-Heterodyne Reception :** M. BOELLA : *Onde Élect.*, 1928, **7**, 500-8.

(See Abs. 1.2.4/18.)

22. **The "Miller" Effect in High Frequency Amplification :** G. H. D'AILLY : *Q.S.T. Franç. Radioélect.*, 1928, **9**, 12-19.

(See Abs. 1.2.7./10.)

23. **Aperiodic H.F. Amplification with Modern Valves :** F. L. DEVEREUX : *Wireless Wld.*, 1929, **24**, 173-74.

(See Abs. 1.3/16.)

24. **A Generalised Analysis of the Triode Valve Network :** F. M. COLEBROOK : *J. Instn. Elect. Eng.*, 1929, **67**, 157-69.

The author gives an analysis of the variation with the resultant anode circuit load of the input impedance and voltage amplification factor of a triode valve. A new mode of presentation is adopted which enables the total range of variation of these quantities to be represented as finite functions of a variable with a limited range of variation.

The most important conclusion is that at very high radio-frequencies the voltage amplification factor may appreciably exceed the voltage factor of the valve, this condition being reached for a finite pure inductive load. Under the same conditions, however, the shunt input impedance is very low.

The most efficient means of associating a tuned circuit with a receiver of relatively low input impedance is considered in an Appendix.

(See also Abs. 2.3.2/33.)

1.2.3. MULTI-STAGE AMPLIFIERS

1. **Vacuum Tubes and their Technical Applications :** H. BARKHAUSEN : *Jahrb. drahtl. Telgr.*, 1919, **14**, 27-47, and **16**, 82-114.

(See Abs. 1.2.2/7.)

2. **The Design of Multi-Stage Amplifiers using Three-Electrode Thermionic Valves :** C. L. FORTESCUE : *J. Instn. Elect., Eng.*, 1920, **58**, 65-74 ; *Radio Rev.*, 1920, **1**, 178-81 ; *Elect. Rev. Lond.*, 1920, **85**, 700-1.

This paper represents a marked advance in the theoretical study of amplifier design, and gives the collected results of experiment. Commencing

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with a single stage of radio-frequency amplification with a resistance or tuned circuit load in the anode circuit, equations are obtained giving the amplification and the input impedance. All four characteristics of the valves are considered and are expressed in a general form as partial differential equations. The equations are then applied to the multi-stage case and some general deductions are obtained.

The problems considered are only those relating to the pure amplifying action, no attempt being made to deal with the action of rectifiers. The theoretical conditions pertaining in a resistance amplifier are examined, and the use of tuned circuits instead of resistances is considered. The input impedance as well as the amplification is examined from a power basis. Retro-action effects are discussed; these are assumed to take place via the anode-grid capacity only. Hence the results obtained cannot be considered to be complete and the analysis is applicable only to medium radio frequencies.

3. **Resistance Amplifiers :** L. BRILLOUIN : *Onde Élect.*, 1922, **1**, 7-17, 101-23.

(See Abs. 1.4.1/2.)

4. **High Frequency Reaction Amplifiers :** L. DUBAR : *Onde Élect.*, 923, **2**, 43-49.

(See Abs. 1.4.1/7.)

5. **High Frequency Amplifiers :** H. T. FRIIS and A. G. JENSEN : *Bell System. Techn. J.*, 1924, **3**, 181-205.

In this paper, a simplified mathematical treatment of the theory of radio-frequency amplifiers is presented and the theory is verified by experiments. This method of mathematical analysis provides a ready means of predicting the performance and action of an amplifier from a knowledge of the fundamental constants of its circuit and places the design of radio-frequency amplifiers on a more rational basis. The paper also includes a description of various methods for quantitatively determining the amplification at radio frequencies. Amongst the arrangements particularly treated are: (1) Tuned circuit and choke-coil amplifiers; (2) loosely coupled transformer amplifiers; (3) amplifiers with several stages and retro-action. The practical aspects of amplification measurements at radio frequencies are also discussed.

The algebraic analysis is extremely elementary and the formulæ used are not derived from first principles. The paper is chiefly of an experimental nature, the method of taking the measurements being described in the measurement section, the theory only being used as a qualitative support of the experimental evidence, no definite agreement being attempted. The experimental results taken at frequencies of 500 to 100 kilo-cycles (600 to 3,000 metres) lead to the following deductions :—

With a given type of amplifier the general shape of the amplification curve is independent of the frequency range at which the amplifier is designed to operate. The other deductions regarding the design of inter-stage radio-frequency transformers are of an elementary character only.

(See also Abs. 4.2/6.)

6. **Valve Amplifiers :** F. OLLENDORFF: *Arch. Elektrotech.*, 1924, **13**, 274-91.

A general system of linear differential equations is shown to be applicable to amplifiers in general. These equations are simplified for audio-frequency amplifiers and are then integrated for quasi-stationary conditions. The

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method of operation is then explained by means of vector and circuit diagrams. For radio frequencies, the integrals of the equations can be obtained easily if quasi-stationary conditions are considered. The integrals in complex form consist of a system of ordinary linear equations, the co-efficients of which can be easily found from the valve data. The design of radio-frequency amplifiers is discussed and an example is given of the design of an amplifier for a frequency of 50 kilo-cycles per sec.

7. **Valve Amplifiers :** H. KAFKA : *Arch. Elektrotech.*, 1925, **14**, 347-54.

This mathematical paper gives an extension of Ollendorff's theoretical analysis of the properties of the valve used as an amplifier.

8. **Resonant Circuit Amplifiers using Four-Electrode Valves :** R. BARTHÉLEMY : *Onde Élect.*, 1927, **6**, 152-60.

(See Abs. 1.2.8/9.)

9. **Resistance Back-Coupling in Valve Amplifiers :** F. SCHIERL : *Arch. Elektrotech.*, 1928, **20**, 346-70.

(See Abs. 1.2.8/13.)

1.2.4. SUPER-REGENERATIVE AND SUPERSONIC-HETERODYNE RECEIVERS

1. **A Study of Heterodyne Amplification by the Electron Relay :** E. H. ARMSTRONG : *Proc. Inst. Radio Eng.*, 1917, **5**, 145-59.

The amplifying action of the oscillating valve receiver is carefully examined. It is found, by separation of the various effects, that there exist three distinct types of amplification. The first, or " equal heterodyne " type, occurs when the local oscillating current is equal to the signal current. The second, or " optimum heterodyne " type, occurs when the local oscillating current is increased to the critical value for maximum response. The third, or " regenerative " type, results from the amplifying action of the valve and its associated circuits. The roughly approximate numerical values of the three types are five, twenty and fifty-fold, making a total amplification of five thousand times or more.

The mechanism of these phenomena is considered in detail with special reference to the limitations of each process.

The conclusions obtained are based entirely upon experiment. The amplification produced in a simple valve circuit was compared with that produced by the same valve when using a self-heterodyne circuit, from which a general idea of the relative amplifications due to the two methods may be obtained.

(See also Abs. 2.3.2/2, 4.3/1.)

2. **A New System of Short-Wave Amplification :** E. H. ARMSTRONG : *Proc. Inst. Radio Eng.*, 1921, **9**, 3-27.

The various methods of amplifying incoming signals at frequencies above 500 kilo-cycles per sec. (600 metres) are described and their limitations considered.

The new method described consists (for continuous wave reception) of the following steps :—

(1) Heterodyning, with the production of a beat frequency which is itself a radio-frequency (for example, 100 kilo-cycles per sec.).
(2) Rectification of the beat signal.
(3) Amplification at the beat radio frequency, preferably by a tuned amplifier.

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(4) Audio-frequency heterodyning of the amplified beat signal.
(5) Rectification of the audio-frequency signal so produced.

For the reception of damped waves or telephony, step (4) is omitted. It is shown theoretically that in this case the quality of the incoming signals should be preserved.

3. **Super-Regenerative Valve Circuits :** E. ARMSTRONG : *Proc. Inst. Radio Eng.*, 1922, **10**, 244-68.

(See Abs. 1.1.2/1.)

4. **On a Type of Oscillation-Hysteresis in a Single Triode Generator :** E. V. APPLETON and B. VAN DER POL : *Phil. Mag.*, 1922, **43**, 177-93.

(See Abs. 1.2.1/11.)

5. **The Bolitho Circuit :** J. B. BOLITHO : *Wireless Wld.*, 1923, **12**, 266-67.

(See Abs. 1.1.2/3.)

6. **A New Circuit :** *Wireless Wld.*, 1923, **12**, 74.

(See Abs. 1.1.2/4.)

7. **On Super-Regeneration :** E. O. HULBURT : *Proc. Inst. Radio Eng.*, 1923, **11**, 391-94.

The action of the Armstrong super-regenerative circuit is given. The writer then offers a simple mathematical analysis which shows that the system amplifies without distortion, and that the amplification increases with signal frequency.

The assumptions made in the theoretical analysis do not represent the actual conditions, which are of a very complex nature, and consequently the simple theoretical discussion given can only approximately explain the observed behaviour of the receiver.

(See also Abs. 1.1.2/5.)

8. **Periodic Trigger Reception :** E. V. APPLETON and F. S. THOMPSON : *J. Inst. Elect. Eng.*, 1924, **62**, 181-86.

An audio-frequency E.M.F. is introduced in the grid circuit of a simple triode oscillator in such a way that radio-frequency oscillations are not produced in the absence of an incoming signal. A small radio-frequency signal is, however, sufficient to produce a train of free oscillations once every audio-frequency cycle. Such free oscillations are applied to a rectifier, and a signal at the impressed audio-frequency is heard in the telephones, the amplitude of the telephone signal being practically independent of the amplitude of the incoming signal.

The chief interest of the present paper lies in the comprehensive analysis of the action of such relay circuits. The analysis is based on the non-linear nature of the triode characteristic.

(See Abs. 1.2.1/11.)

9. **An Oscillographic Study of Super-reaction :** DAVID, DUFOUR and MESNY : *Onde Élect.*, 1925, **4**, 175-200.

The paper gives a review of the principles of Armstrong's method of reception by " super-regeneration," and describes the experimental arrangements, employing a Cathode ray oscillograph, for investigating (1) the special state of sensitivity, in which the receiver oscillates violently under the influence

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of a small impulse ; (2) the proportionality of the oscillations to the applied E.M.F. ; (3) the variation of amplitude with the frequency ; and (4) the action of the local oscillator. A number of oscillograms are given showing the variations in grid and anode voltages for radio and audio-frequencies, with varying conditions of filament heating, in the presence and absence of a maintained transmission and with and without a local oscillator ; the results are examined in some detail. A mathematical statement of the growth and decay of the oscillations is given, curves being deduced which are used to explain the effects of the various influences mentioned above.

10. **An Analysis of Regenerative Amplification :** V. D. LANDON and K. W. JARVIS : *Proc. Inst. Radio Eng.*, 1925, **13**, 709-53.

(See Abs. 1.4.1/15.)

11. **Use of Double-grid Valves in Frequency Changing Circuits :** J. DE MARE, R. BARTHÉLEMY, H. DE BELLESCIZE and L. LEVY : *Onde Élect.*, 1926, **5**, 150-80.

Under this heading are included the four papers contributed by the above authors, as they deal with the same subject.

Double-grid valves can be used as the frequency-changing valve in the supersonic-heterodyne type of circuit, the inner grid being supplied from the local oscillator and the outer grid being connected to the aerial system.

The exact nature of the effect obtained is a matter of considerable controversy. The first three authors hold the view that the action in the double-grid valve is of a different nature from that in the ordinary heterodyne circuit and is independent of any curvature in the valve characteristics. LEVY, however, takes the opposite view and maintains that there must be curvature of the characteristic and consequent detection, otherwise frequency-changing would not occur. The nature of this effect is discussed in Abs. 2.3.3/12, 14, 15, 16.

12. **Valve Frequency Changers :** R. BARTHÉLEMY : *Rev. Gén. Elect.*, 1926, **19**, 663-70.

In the first part of the paper the author deals with the theory of the frequency changer as used in the usual supersonic-heterodyne circuit in which the local oscillations are applied to the anode circuit of a triode, the grid circuit of which is supplied with the signal oscillations. The second portion of the paper deals with the case in which a double-grid tetrode is used. In this case, the outer grid is connected to the aerial circuit and thus receives the incoming signals ; the inner grid contains in its connections an oscillatory circuit so coupled to a suitable coil in the anode circuit as to generate the necessary local oscillations. There is also in the anode lead the usual circuit tuned to the beat or intermediate frequency. Applying the theory with suitable extensions developed in the first part of the paper, the author shows that the mechanism by which the intermediate frequency is produced is essentially different from that of the usual supersonic-heterodyne principle. In particular it is shown that the amplitudes of the beats forming the intermediate frequency is independent of the beating frequencies, and, further, that the power available in the intermediate-frequency circuit is limited only by the power of the local oscillator.

(See also Abs. 2.3.3/17.)

13. **Supersonic Transformers :** N. W. MCLACHLAN : *Wireless Wld.*, 1926, **19**, 631-34, 680-84, 715-18, 777-78 ; *Wireless Wld.*, 1927, **20**, 21-24.

(See Abs. 1.3/7.)

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14. **Relaxation Oscillations :** B. VAN DER POL : *Jahrb. drahtl. Telegr.*, 1926, **28**, 178-84 and *Phil. Mag.*, 1926, **2**, 978-92.

(See Abs. 1.2.1/20.)

15. **Super-heterodyne Receivers :** E. H. ULLRICH and A. H. REEVES ; *Experimental Wireless*, 1927, **4**, 652-56.

A mathematical treatment is given which explains the phenomenon stated in Abs. 1.1.3/8, and a circuit is shown which avoids this coupling to such an extent that the variation of grid filament volts at the beating oscillator frequency with change of the input circuit tuning is reduced from 300 per cent. to less than one-half of 1 per cent.

(See also Abs. 1.1.3/8.)

16. **The Screened Valve in Audio-Frequency Circuits :** N. W. MCLACHLAN : *Wireless Wld.*, 1927, **21**, 536-38, 577-79.

(See Abs. 3.2.7/2.)

17. **Measurements on Radio-Frequency Amplifiers :** R. L. SMITH, ROSE and H. A. THOMAS : *Wireless Wld.*, 1927, **20**, 131-35, 171-77, 199-202-237-39.

(See Abs. 4.2/12.)

18. **The Calculation of Medium Frequency Amplifiers for Super-heterodyne Reception :** M. BOELLA : *Onde. Élect.*, 1928, **7**, 500-8.

It is shown how one can determine, in a transformer coupled stage with tuned secondary, the primary inductance to give maximum amplification, provided the constants of the secondary are known. From this are deduced the elements of a medium frequency amplifier for a supersonic-heterodyne receiver, to receive telephony under good conditions of selectivity and fidelity. The example taken has three stages, plus the rectifier, for a carrier wave of frequency 200 kilo-cycles per sec. (1,500 metres) to give practically constant amplification for modulation frequencies from 100 to 4,000 p.p.s.

19. **A Note on Some Interfering Oscillations Experienced in a Supersonic-Heterodyne Receiver :** R. L. SMITH-ROSE : *Experimental Wireless*, 1928, **5**, 673-76.

(See Abs. 1.2.10/13.)

20. **Transmission and Reception by Super-Regenerative Receivers :** G. BEAUVAIS : *Onde Élect.*, 1928, **7**, 206-9.

(See Abs. 1.2.10/14.)

21. **Super-Regenerative Circuits :** P. DAVID : *Onde Élect.*, 1928, **7**, 217-59.

A very comprehensive paper on the above subject, in which the theory is given largely in a non-mathematical form, the author commenting on the practical impossibility of a rigorous treatment which would be capable of any useful physical interpretation. Three main types of super-regeneration are distinguished, any one of which may be merged almost imperceptibly into the others : type A, or " telephonic " super-regeneration, which is the type originally due to Armstrong ; type B, or " stroboscopic " super-regeneration ; and type C, or " anti-disturbance " super-regeneration. The conditions appropriate to the three types are stated, and the theory is given with an experimental verification. The article concludes with a bibliography of some twenty-three references to the subject.

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22. **Present Developments in Super-heterodyne Receivers :** G. L. BEERS and W. L. CARLSON : *Proc. Inst. Radio Eng.*, 1929, **17**, 501-15.

(See Abs. 1.1.3/14.)

23. **The Problem of the most Economic Multiple Frequency-changing in Reception :** F. AIGNER : *Jahrb. drahtl. Telegr.*, 1929, **33**, 9-15.

The paper begins with a comparison between neutrodyne and frequency-changing methods, and points out the "erroneous idea" that a super-heterodyne is, on theoretical grounds, more selective than a neutrodyne, and shows that no such superiority exists in theory. On practical grounds, however, the frequency-changing method possesses an advantage, and the author's first object is to put this on the most economic basis by reducing the number of fixed local oscillators to a minimum. His second object, dealt with in the part of his paper still to appear, is to develop an apparatus, for un-modulated and modulated waves, which shall retain all the advantages of the method and possess as well advantages attainable with no existent apparatus, particularly with regard to selectivity. The present paper deals with apparatus with normal selectivity ; only two local oscillators, or one yielding two harmonics, are found to be necessary ; or one oscillator giving a fundamental and harmonics can, by a combination of difference frequencies with these harmonics, yield a whole series of suitable oscillations.

1.2.5. RADIO-FREQUENCY FILTER CIRCUITS. (See 1.1.9.)

1.2.6. VALVE NOISES AND THE "SCHROT" EFFECT

1. **Electric Oscillations :** W. SCHOTTKY : *Ann. Physik.*, 1918, **57**, 541-67.

A theoretical discussion of the phenomenon associated with electrical oscillations of small amplitude, with particular reference to sustained oscillations in vacuum tubes with two or three electrodes due to the evaporation of successive electrons and the consequent rapid succession of impulses in the tuned anode circuit.

2. **Determination of the Elementary Electrical Quantum from the "Schrot Effect."** : C. A. HARTMANN : *Ann. Physik.*, 1921, **65**, 51-78.

This paper is concerned with the application of the thermionic amplifier to the determination of small electrical variations such as occur in ionisation currents or disintegration currents of radio-active material. The object is the measurement of the spontaneous current variations in high-vacuum discharge tubes, a subject which has been previously dealt with theoretically by SCHOTTKY, and called by him the "Schrot" effect. SCHOTTKY has advanced a formula

$$I = \sqrt{(2\pi)}\ \sqrt{(\epsilon i_0/_i)}$$

for the effective mean value I of the current, where ϵ is the elementary quantum charge. In terms of voltage variation V and impedance Z of the compared circuit,

$$I = \frac{V}{Z_0} \text{ and } \epsilon = \frac{V^2}{i_0 \omega_0 Z_0^2}$$

where $\omega_0 = \frac{2\pi}{\epsilon}$

The particular type of two-electrode valve used in the experiments is then illustrated and described in great detail, as also the bridge and telephone,

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and the oscillatory circuits used to make the comparison by aural means. A further apparatus for directly determining the value of ϵ is also illustrated and described, together with the theory of the methods used.

The author summarises his results as follows :—

(1) The quantum effect is greater than if due to purely thermal effects and diminishes with increase of frequency.

(2) Investigation shows that the effect is not of external origin and (*a*) the internal resistance of the valve is independent of the frequency but very dependent on the saturation current, and the valve possesses capacity which increases with the saturation current ; (*b*) the damping resistance of the oscillatory circuit depends largely upon the frequency.

(3) Exact measurements show the function e (ω_0) to have limiting values which are not due to errors of measurement.

(4) It is demonstrated that the apparent smallness of the quantum effect is a function of the saturation currents.

(5) It is advanced as a qualitative explanation of the effect, that it is due to the thermal conductivity of the filament and the influence of the emitted electrons upon the surface temperature of the filament. The Schottky assumption of its being an independent effect does not hold for the particular saturation current used. The apparent valve capacity has, likewise, to be attributed to thermal conductivity.

(See also Abs. 4.1.2/1.)

3. **Determination of the Elementary Electrical Quantum from the "Schrot Effect."** : J. B. JOHNSON : *Ann. Physik.*, 1922, **67**, 154-56.

SCHOTTKY and HARTMANN have published papers on this subject. It is here pointed out that SCHOTTKY has apparently made a mistake in the evaluation of a certain integral, with the result that calculation of the elementary electrical quantum ϵ from Hartmann's observations gives a value 500 to 1,000 times too small. Using the corrected value of the integral, the value of ϵ is found to be of the right order of magnitude. At radio frequencies, the value of ϵ is almost correct, while at audio-frequencies, the value is not more than six times too great.

4. **Calculation and Discussion of the "Schrot Effect"** : W. SCHOTTKY : *Ann. Physik.*, 1922, **68**, 157-76.

JOHNSON has pointed out an error in Schottky's previous paper. This paper gives a new method of evaluating the integral concerned. This results in the introduction of a second correction factor of Z, which brings the calculated value for the elementary electronic charge to the right order of magnitude, the mean value of all the experimental results being very near the usually accepted value of $1{\cdot}56 \times 10^{-19}$ coulomb.

5. **The Schottky Effect in Low Frequency Circuits** : J. B. JOHNSON : *Phys. Rev.*, 1925, **26**, 71-85.

(1) In the absence of space charge this effect, discovered by SCHOTTKY, which depends on the probability fluctuations of electron emission from a filament, has been measured over a considerable range of conditions in resonant circuits of which the natural frequency was varied from 8 to nearly 6,000 p.p.s. The effect is much larger in the lower range of frequencies than the theory predicts. With a tungsten filament, the ratio of observed to theoretical effect e^1/e is about 0·7 for frequencies above 200, but increases rapidly to 50 at 10 cycles per second. With an oxide-coated filament the ratio increases from 1 at 5,000 cycles to 100 at 100 cycles. This is interpreted to mean that the emission of electrons is not strictly chaotic, but is influenced by irregular temporal changes in the cathode emissivity. In a radio-frequency circuit these changes become imperceptible, and the emission is effectively at random.

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(2) When the current is limited by the space charge, the Schottky effect decreases because of the inter-action of the electrons, and other disturbances may act upon the space charge so as to completely mask the remanent Schottky effect. The magnitude of the disturbances in amplifying vacuum tubes cannot, therefore, be predicted from measurements on the true Schottky effect.

6. **The Theory of the Schroteffekt** : T. C. FRY : *J. Franklin Inst.*, 1925, **199**, 203-20.

The general theory of the Schrot effect is reviewed and it is shown how its value can be measured by the difference between the energy actually dissipated in a tuned system in the valve network, and the energy that would be dissipated if the electron stream were a uniform flow.

The experimental results agree fairly well with the theory, but at low frequencies the discrepancy is so great that a modification of Schottky's theory seems desirable.

7. **The Theory of the " Schrot-effect "** : N. CAMPBELL : *Phil. Mag.*, 1925, **50**, 81-86.

Attention is drawn to a general proposition relating the fluctuations in any fortuitous phenomena and those in an instrument by which it is observed. Use of it would avoid the elaborate calculations upon which the study of the Schrot effect has been based, and enable the observations to be interpreted much more directly in terms of measurements.

The results of HULL and WILLIAMS are discussed and it is shown that conclusions other than those that they draw are not excluded.

8. **Small-Shot Effect and Flicker Effect** : W. SCHOTTKY : *Phys. Rev.*, 1926, **28**, 74-103.

JOHNSON observed, under certain conditions, voltage fluctuations across connected resonant circuits which were much larger than the theory of the small-shot effect would lead one to expect. Analysing Johnson's curves, it is found that this effect increases as the square of the electron current i_0 instead of as the first power, as in the case of the small-shot effect. The trend with the natural frequency of the connected circuit is likewise different from that observed in the small-shot effect. The elementary atomic process underlying the flicker effect is the appearance of an individual foreign atom or molecule in the surface of the cathode, changing the ability of the surface to emit electrons so long as the foreign atom remains. The influence exerted upon the current by foreign atoms in the surface may be calculated with sufficient approximation from the electrical image theory and Langmuir's doublet theory. From the assumptions that the elementary atomic processes are independent of one another, and that the length of stay of the foreign atoms in the surface conforms to statistical laws, the frequency distribution of the flicker effect is derived and compared with Johnson's experiments, whence it is possible to determine the length of stay of the foreign atom in the surface. From calculations, it is inferred that the essential cause of the flicker effect with oxide cathodes lies in a continuous exchange of positions by the components of the oxide.

The variation of the flicker effect with the square of the total current follows from the assumption that the current change due to one foreign atom is proportional to the total current. From this assumption also follows the relative independence upon temperature. A value is given for the mean square voltage fluctuation produced by the effect in terms of the impedance of the connected circuit and formulæ are given for circuits of various kinds. Measurements with two circuits, one being resonant and the other a pure resistance, should provide a definite check on the calculated length of stay of the foreign atoms.

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9. **Theory of the "Shot" Effect :** H. A. WHEELER : *Phys. Rev.*, 1927, **29**, 903.

Previous derivations of the magnitude of the "Schrot" effect have been based on equations derived in the abstract by the theory of probability. In this paper, a simple derivation of the equation

$$(I_0^2)_{\text{mean}} = \frac{eI_a}{2RC}$$

(where I_a is the average space current) is given in terms of the familiar discharge current in a simple series circuit (R, L, C). This is followed by a Fourier integral derivation of the continuous frequency spectrum of the current fluctuations :

$$\frac{d(I^2)_m}{d\omega} = \frac{eI_a}{\pi}$$

The simple derivation is thereby linked with former work and the correctness of the assumptions justified.

10. **Thermal Agitation of Electricity in Conductors :** J. B. JOHNSON : *Phys. Rev.*, 1928, **32**, 97-109.

A statistical fluctuation of electric charge exists in all conductors, producing random variation of potential between the ends of the conductor. The effect of these fluctuations has been measured by an amplifier and thermo-couple, and can be expressed by the formula $I^{-2} = \left(\frac{2kT}{\pi}\right) \int_0^\infty R(\omega)/Y(\omega)/^2 d\omega$. I is the observed current in the thermo-couple, k is Boltzmann's gas constant, T is the absolute temperature of the conductor, R (ω) is the real component of impedance of the conductor, Y (ω) is the transfer impedance of the amplifier, and $\omega/2\pi \times f$ represents frequency. The value of Boltzmann's constant obtained from the measurements lie near the accepted value of this constant. The technical aspects of the disturbance are discussed. In an amplifier having a range of 5 kilo-cycles per second, and an input resistance R, the power equivalent of the effect is $V^{-2}/R = 0{\cdot}8 \times 10^{-15}$ watt, with corresponding power for other ranges of frequency. The least amount of valve noise is equivalent to a resistance $R_e = 1{\cdot}5 \times 10^5\, i_b/\mu$ where i_b is the space current in milli-amperes and μ is the effective amplification of the valve.

1.2.7. INPUT IMPEDANCE

1. **The Audion as a Circuit Element :** H. W. NICHOLS : *Phys. Rev.*, 1919, **13**, 404-14.

The effects of inter-electrode capacities are examined and applied to some examples of amplifiers and oscillators. The equations giving the input impedance of the valve network are deduced by Kirchoff's Laws, assuming that the static characteristics are linear.

It should be noted that the results are similar to those of the classical work of MILLER and are prior to his work : the results, however, are not presented so satisfactorily.

(See also Abs. 1.2.2/9.)

2. **Dependence of the Input Impedance of a Three-Electrode Vacuum Tube upon the Load in the Plate Circuit :** J. M. MILLER : *Sci. Pap. Bur. Stand. Wash.*, 1919, 15, No. 351, 367-85.

In this important paper the valve circuit network is carefully examined, formulae of a practical form being given for the calculation of the input

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capacity and resistance, here considered to be in series. Satisfactory experimental confirmation is given. It is shown that a resistance or capacity anode circuit load produces a positive input resistance, and an inductive load a negative resistance.

(See also Abs. 4.1.3/1.)

3. **The Design of Multi-Stage Amplifiers using Three-Electrode Thermionic Valves :** C. L. FORTESCUE : *J. Inst. Elect. Eng.*, 1920, **58**, 65-74 ; *Radio Rev.*, 1920, **1**, 178-81 ; *Elect. Rev. Lond.*, 1920, **85**, 700-1.

(See Abs. 1.2.3/2.)

4. **Input Impedance of the Thermionic Amplifier:** S. BALLANTINE : *Phys. Rev.*, 1920, **15**, 409-20.

At audio-frequencies, the input circuit behaves as a pure capacity; at radio-frequencies, due to the inherent capacity of the anode and grid electrodes, the input impedance resembles the combination of a condenser shunted by a resistance, this being dependent upon the load in the anode circuit. A resistive load gives a positive input resistance and represents a power loss. An inductive load above a certain critical value manifests itself as a negative shunt resistance and thus supplies energy from the anode to the grid circuit. These feed-back effects are occasioned solely by the grid to anode capacity and depend upon its magnitude. The reason for self-oscillation in such circuits is obvious, and occurs when the fed-back energy is greater than the input circuit losses. Experimental data is given confirming the theory.

5. **Grid-filament Conductivity :** F. M. COLEBROOK : *Electrician*, 1923, **91**, 574-75.

In this paper the necessity for maintaining the grid at a sufficiently negative potential to prevent grid current is illustrated by vector analysis and a practical example.

The grid-filament conductivity considered is not that due to the anode circuit load and the anode-grid inter-electrode capacity.

6. **The Performance of Amplifiers :** H. A. THOMAS : *J. Instn. Elect. Eng.*, 1926, **64**, 253-68, *Disc.* 268-78 ; *Experimental Wireless*, 1926, **3**, 15-18 ; *Elect. Rev. Lond.*, 1926, **98**, 73-74.

The portion of this paper relevant to this section gives a graphical theoretical method of predicting the effect of a radio frequency amplifier on the tuned input circuit. The increase in effective resistance of this input circuit due to the presence of a stable radio-frequency amplifier is calculated and experimental results are given in confirmation of the theory.

(See Abs. 1.4.1/16, 3.4/9, 4.2/8, 4.4.1/7, 4.5/7, 4.6/1.)

7. **On the Values and the Effects of Stray Capacities in Resistance-Coupled Amplifiers :** M. v. ARDENNE and W. STOFF : *Proc. Inst. Radio Eng.*, 1927, **15**, 895-901.

(See Abs. 3.2.2/12.)

8. **Calculation of Apparent Capacity in Resistance Amplifiers :** M. v. ARDENNE and W. STOFF : *Jahrb. drahtl. Telegr.*, 1927, **30**, 86-89.

A formula for the calculation of the apparent capacity of an amplifier stage with any anode load is given, in which the phase displacement between E_a and E_g must be taken into account. The retro-action of the anode load on the grid circuit in the case of the individual stages of a three-stage resistance-capacity-coupled amplifier is examined by means of a numerical example.

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This retro-action needs for its correct representation an ohmic resistance as well as an additional capacity to be added to the grid circuit. It is further shown that only the first stages of a three-stage amplifier possess any perceptible frequency distortion, due to the large loading effect of the following stages.

9. **The Input Impedance of Thermionic Valves at Low Frequencies:** L. HARTSHORN: *Proc. Phys. Soc. Lond.*, 1927, **39**, 108-23.

This paper points out, as a result of careful bridge measurements on triodes, that the inter-electrode impedances are not pure capacities but consist of a capacity in combination with a resistance, i.e., the power-factor of the inter-electrode capacities is not zero. The input circuit is regarded as being equivalent to a condenser with a definite phase angle ϕ or "loss angle" $\delta = 90° - \phi$, and the results are expressed by stating the effective capacity and value of tan δ for each set of experimental conditions. A series of measurements made on an R valve are recorded, and it is shown that the results are in good agreement with the theoretical investigations of MILLER and NICHOLS. The theoretical investigation has been extended to allow for the effect of dielectric losses in the valve, since these were found by experiment to be rather large and to have an appreciable effect on the capacity and phase angle of the input circuit. It is shown that the effective capacity may vary from about 10 $\mu\mu$ F. to 100 $\mu\mu$ F. for an R valve, and the phase angle may vary from about 80° leading to 126° leading, depending mainly on the load in the anode circuit. Values of phase angle greater than 90° correspond to a negative resistance or negative power factor, and occur when the load in the anode circuit is inductive. The variations of input capacity and phase angle with filament voltage, anode voltage, input voltage and frequency are also investigated.

The valve network is examined by determinant mathematics in a more comprehensive manner than in the analysis due to MILLER, the paper representing the latest attempt to examine theoretically the input impedance of a stage of amplification.

10. **The "Miller" Effect in High Frequency Amplification:** G. H. D'AILLY: *Q.S.T. Franç. Radioélect.*, 1928, 12-19.

The author disapproves of the mathematical treatment which MILLER has given to the phenomenon which bears his name, and proceeds to give his own analysis of the effect.

11. **Internal Capacity of Multi-Electrode Valves:** C. RAJSKI: *Onde Élect.*, 1928, **7**, 461-74.

The inter-electrode capacities of a valve cannot be determined theoretically purely from electrostatic considerations, since the space-charge prevents the problem from being one of simple electrostatics. The author also concludes that the usual three capacities—grid to filament, grid to anode, anode to filament—are not sufficient, and proposes four capacities, viz., true grid, true anode, anode to grid, and grid to anode, the two latter being, in general, not equal. These capacities are found mathematically for both plane and cylindrical electrodes.

12. **A Generalised Analysis of the Triode Valve Network:** F. M. COLEBROOK: *J. Instn. Elect. Eng.*, 1929, **67**, 157-69.

(See Abs. 1.2.2/24, 2.3.2/33.)

1.2.8. AMPLIFIERS USING SPECIAL VALVES

1. **The Four-Electrode Valve:** N. NOZIÈRES and P. GIROUD: *Onde Élect.*, 1924, **3**, 583-90.

(See Abs. 1.1.8/2, 2.3.3/3.)

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2. **Use of Double-grid Valves in Frequency-Changing Circuits :** J. DE. MARE, R. BARTHÉLEMY, H. DE BELLESCIZE and L. LEVY : *Onde Élect.*, 1926, **5**, 150-80.

(See Abs. 1.2.4/11, 2.3.3/13, 14, 15, 16.)

3. **Characteristics of Shielded-Grid Pliotrons :** A. W. HULL and N. H. WILLIAMS : *Phys. Rev.*, 1926, **27**, 432-38.

(See Abs. 1.1.8/9.)

4. **Valve Frequency Changers :** R. BARTHÉLEMY : *Rev. Gén. Elect.*, 1926, **19**, 663-70.

(See Abs. 1.2.4/12, 2.3.3/17.)

5. **Employment of Four-Electrode Valves :** B. DECAUX : *Onde Élect.*, 1927, **6**, 1-18.

A review of the general properties is given, and it is shown how these valves can be arranged to give a very large amplification or to use a very low anode potential. Mention is made of the negative resistance phenomena and some interesting circuits arising from it.

(See also Abs. 2.3.3/19.)

6. **The Shielded Plate Valve as a High-Frequency Amplifier :** R. T. BEATTY : *Experimental Wireless*, 1927, **4**, 619-25.

The general properties of the theoretically perfect screen-grid valve are enunciated, together with a description of Hull's valve.

In this valve a fourth electrode or shield made of wire gauze envelops the anode as completely as possible and is kept at a fixed positive potential. The properties of a perfect valve of this type are that the anode-grid capacity is zero and that the differential resistance of the valve is infinite. In the screen-grid valve due to HULL, these quantities are $0 \cdot 006$ $\mu\mu$ F. and 5×10^5 ohms. respectively. The voltage amplification factor obtainable at a frequency of 10,000 kilo-cycles per sec. is 7. In commercial screen-grid valves, the anode-grid capacity is $0 \cdot 1$ $\mu\mu$ F. and the resistance is $1 \cdot 25 \times 10^5$ ohms ; the voltage amplification ranges from 13 at a frequency of 4,000 kilo-cycles per sec. to 58 at 500 kilo-cycles per sec., and the total amplification from 990 to 4,100 for these values respectively.

The stability of one stage is theoretically examined.

7. **Screened Valves :** P. W. WILLANS : *Wireless Wld.*, 1927, **21**, 107-10.

The elementary principle of the screen-grid valve is given, the valves of HULL being described. A general circuit layout of a four-valve receiver is described in detail.

8. **The New Screened Valve :** N. W. MCLACHLAN : *Wireless Wld.*, 1927, **21**, 260-63, 307-10.

A general examination of the performance of a screen-grid valve is given and the characteristics are shown. Its use as a radio-frequency amplifier is considered and a simple summary of its properties forms the conclusion.

9. **Resonant Circuit Amplifiers using Four-Electrode Valves :** R. BARTHÉLEMY : *Onde Élect.*, 1927, **6**, 152-60.

The author shows how self-oscillation in a multi-stage amplifier can be prevented by the use of four-electrode valves. The selectivity can be materially improved and it is claimed that six stages of radio-frequency amplification at a frequency of 1,000 kilo-cycles per sec. (300 metres) can be made stable by their use.

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10. **Characteristic Curves of the Four-Electrode Valve :** N. R. HALL : *Experimental Wireless,* 1928, **5,** 198-200.

An elementary paper giving experimental characteristic curves only.

11. **Approximate Theory of the Screened-Grid Valve :** B. C. BRAIN : *Experimental Wireless,* 1928, **5,** 179-83.

The author develops a simple theory of the valve and shows how the general shape of the characteristic and the valve constants (amplification factor, mutual conductance, and inter-electrode capacity), can be predicted from a knowledge of the structural constants of the electrodes.

12. **Good Quality in High-Frequency Amplifiers :** C. C. INGLIS : *Experimental Wireless,* 1928, **5,** 132-34.

The author gives a formula from which the width of a resonance curve may be obtained for any fraction of its height. Using this formula and postulating a 5 per cent. reduction in amplitude as that permissible for good quality, he shows that a screen-grid valve with the ordinary low-loss coil will cause bad quality owing to side-band cut-off.

The Editor of *Experimental Wireless* points out that the criterion adopted (only a 5 per cent. reduction in amplitude) is in his opinion far too severe.

13. **Resistance Back-coupling in Valve Amplifiers :** F. SCHIERL : *Arch. Elektrotech.,* 1928, **20,** 346-70.

The author states theoretically the behaviour of two resistance-coupled valves with a pure resistance retro-action (the kallirotron). In a valve containing two grids, the current to the space-charge grid can be utilised to give the same effect. The author shows that with the increase in the retro-action resistance, the steepness of the amplifying characteristic can be increased indefinitely, and that instability arises from slight irregularities in the characteristic curves.

14. **Some Characteristics and Applications of Four-Electrode Tubes :** J. C. WARNER : *Proc. Inst. Radio Eng.,* 1928, **16,** 424-46.

(See Abs. 1.1.8/16.)

15. **Double-Grid Valves :** J. DANTSCHER : *Arch. Elektrotech.,* 1928, **20,** 333-45.

Comprehensive theoretical and graphical studies are given of the double-grid valve derived from the static characteristics.

1.2.9. SCREENING OF RADIO-FREQUENCY AMPLIFIERS

1. **On the Electro-magnetic Screening of a Triode Oscillator :** R. L. SMITH-ROSE : *Proc. Phys. Soc. Lond.,* 1922, **34,** 127-38.

The results of the many experiments on screening described in the original paper show that the most complete method of screening a valve receiver is to enclose it in a box made of metal of suitable thickness for the frequency used, and to seal hermetically all joints in the box and its lid. The smallest crack or hole is sufficient to allow of the escape of a detectable amount of the radio-frequency energy, and it is shown that iron is far more effective than copper of the same thickness in preventing direct penetration of radio-frequency magnetic fields through the metal.

Descriptions are given of some practical arrangements adopted to screen radio-frequency apparatus.

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2. **Note on Electro-magnetic Screening :** R. A. W. WATT and J. F. HERD : *Wireless Wld.*, 1923, **11**, 532-34.

The paper gives experimental results on the screening effect of sheet iron. The signal strength obtained from a local and a distant station by the use of a frame-aerial was roughly measured, with and without an iron amplifier screen, from which it was deduced that about 40 per cent. of the output energy in the unscreened case was due to direct reception of the amplifier, batteries, leads and telephones. It is concluded that, in all frame aerial work, particular care must be taken to ensure that the amplifier itself does not behave as an effective antenna.

3. **Some experiments on the Screening of Radio Receiving Apparatus :** R. H. BARFIELD : *J. Instn. Elect. Eng.*, 1924, **62**, 249-64.

An investigation carried out under the direction of the RADIO RESEARCH BOARD to determine the effect of various forms of metallic screen on a frame receiver, the wave-lengths used being large compared with the dimensions of the screen. The method used was to connect two receiving loops in circuit so as to oppose one another when acted on by a given radio wave. The coils are spaced so that there is no mutual action, and they are then adjusted until there is no signal. One coil is then placed inside the screen, and the other is rotated until the signal again disappears. A knowledge of the angles made by the two coils, with the direction of the incoming wave, enables the ratios of the intensities outside and inside the screen to be determined. In this manner the effects of a very large number of different types of screens were investigated.

It is shown that the electric and magnetic fields may be treated separately ; arrangements have been devised for screening a receiving apparatus from either the electric or the magnetic fields. It is very difficult, however, to get perfect screening, even the smallest gaps allowing a detectable amount of energy to enter. On the other hand, such a coarse material as a 1-inch mesh wire netting will give a screening effect of the order of 95 per cent.

The paper is a very comprehensive one and deals with the matter both experimentally and theoretically.

4. **Shielding of Electric and Magnetic Fields :** J. H. MORECROFT and A. TURNER : *Proc. Inst. Radio Eng.*, 1925, **13**, 477-505.

An account is given of an investigation, partly experimental and partly theoretical, of the shielding of electric and magnetic fields for both constant and changing fields. The use of iron shells or sheets for shielding against the fields of permanent magnets, as well as those set up by electric currents, is considered, and the best form for the iron sheets is given for various cases. A given length of iron appears to shield better when flat than when partly surrounding a conductor. Magnetic and electric leakage is discussed and examples are given of circuits with no external field. An expression is obtained for the shielding effect of a short-circuited coil. This was verified experimentally. The shielding is perfect for zero resistance and is zero for zero inductance. As the inductance of the coil increases, the resistance and the frequency of the alternating magnetic field remaining constant, the shielding increases, approaching a limit. The shielding effect of metal sheets against changing magnetic fields is also considered. Other things being constant, the shielding effect increases as the frequency of the magnetic field increases. The shielding also increases with the thickness of the metal, but decreases with increased resistivity of the material. Shielding is always accompanied by an increase in the effective resistance of the shielded circuit. For a given metal at a given frequency, a certain thickness introduces a maximum resistance into the shielded circuit. The thickness giving maximum added resistance decreases as the frequency increases. The shielding qualities of a sheet of metal are decreased by slits across the effective shielding area

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of the metal. The shielding of copper mesh depends on the connections between the wires of the mesh, so that good connections give good shielding, while poor connections give poor shielding. The results for a permeable metal show that the shielding may be better or worse than that afforded by copper, depending on the frequency of the alternating magnetic field and the thickness and resistivity of the material.

5. **How to Obtain Effective Screening :** R. L. SMITH-ROSE : *Wireless Wld.*, 1925, **17**, 694-99.

The theory and action of screens is simply described. Certain practical illustrations are then given showing what type of screen is required for various purpcses. For the screening of comparatively large volumes, ordinary wire-netting of about 1-inch mesh is fairly effective and at the same time convenient to apply. Small instruments are best screened by being enclosed in boxes made of tinned iron sheet. Iron is better than copper. All joints must make good metallic contact.

6. **Shielded Neutrodyne Receiver :** J. F. DREYER and R. H. MANSON : *Proc. Inst. Radio Eng.*, 1926, **14**, 217-47, 395-412.

(See Abs. 1.1.4/6.)

7. **Disc. on " The Shielded Neutrodyne Receiver "** by DREYER and MANSON : L. A. HAZELTINE : *Proc. Inst. Radio Eng.*, 1926, **14**, 395-412.

(See Abs. 1.1.4/6.)

8. **A General Theorem on Screened Impedances :** R. M. WILMOTTE : *Phil. Mag.*, 1928, **6**, 788-95.

The theorem leads to the fact that the impedance of any piece of screened apparatus can be given without stating which terminal should be kept at the potential of the screen, provided that one of the two terminals is so kept at the potential of the screen, but without connection to it. Certain limitations are discussed.

1.2.10. INTERFERENCE AT RADIO-FREQUENCIES

1. **A System for Measuring the Amount of Static:** A. M. CURTIS : *Proc. Inst. Radio Eng.*, 1921, **9**, 225-27.

The author describes an arrangement for measuring the approximate amount of interference received. An audio-frequency transformer is connected in series with the telephones, and the atmospherics, converted into audio-frequency waves by the rectifier, are passed into a rectifying valve. The resultant D.C. pulses are used to charge a very large condenser, which is discharged slowly through a galvanometer in series with a very high resistance. If the condenser and the resistance are large enough, the deflection of the galvanometer will be reasonably steady, and its deflection will depend upon the average intensity and frequency of occurrence of the atmospherics. Such a system could be standardised and used for a comparison of the amount of atmospheric interference received at different stations.

The arrangement is extremely simple and would appear to be very practicable.

2. **Retro-action in Tuned Amplifiers :** H. DE BELLESCIZE : *Radio-électricité*, 1922, **3**, 183-87, 238-42.

(See Abs. 1.4.1/6.)

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3. Radio Receiving Equipment : F. CONRAD : *Proc. Inst. Radio Eng.*, 1922, **10**, 426-39.

Some of the possibilities in the direction of eliminating interference by various circuit arrangements are discussed.

(See also Abs. 1.4.1/4.)

4. Reduction of Interference in Receiving Sets : E. W. MARCHANT : *Wireless Wld.*, 1923, **12**, 426-31, 463-65, 496-98.

After referring to the work of APPLETON and WATT on the wave-shape of atmospherics, the author describes various methods of reducing interference. These include the use of balanced valves and crystals, various arrangements devised by WEAGENT, retro-action devices, Hinton rejector circuits, tuned transformers and tuned anode circuits.

5. Signal-to-Static-Interference Ratio in Radio Telephony : J. R. CARSON : *Proc. Inst. Radio Eng.*, 1923, **11**, 271-74.

Several general propositions are stated relative to the signal-to-atmospheric ratio, in single and double side-band transmission, indicating a superiority in practice for the former system.

The paper is non-mathematical.

6. Effect of Disturbances on Reaction Type Receivers : L. BRILLOUIN and E. FROMY : *Onde Élect.*, 1924, **3**, 181-89.

The paper examines the effect of a powerful interfering C.W. signal upon the sensitivity of a receiver tuned to a weak neighbouring signal. Experiments were carried out, using two heterodynes, one of small power representing the station to be received, and the other of large power representing the interfering station. The reception of the weak heterodyne was examined with various degrees of the interfering signal. The experiments indicated that, in general, the interference reduced the sensitivity to a weak signal, due to a disturbance of the normal retro-action. It is also shown that, if a high anode voltage is used, bringing the upper part of the anode-current characteristic into use, that the interference masks the received signal completely and starts oscillations in the receiver.

A mathematical treatment is given (i) for a single valve receiver, and (ii) for a multi-valve receiver, employing retro-action from the anode circuit of the last valve. The principal causes of trouble are found to be related to the phenomenon of rectification, particularly grid-circuit rectification, and the intensity of the perturbation is found to vary as the square of the amplification of the $(n - 1)$ first stages of the radio-frequency amplifier.

Hence, to reduce the effect of such interference, it is necessary—

(i) to reduce to a minimum the amplification before retro-action ;
(ii) to eliminate rectification in the first valve, by suppressing all resistance in the anode and grid circuits and by using an anode voltage sufficiently high to bring the functioning point on to the straight part of the characteristic ;
(iii) to avoid multiple rectification.

If such precautions are taken, the effects of interference consist solely of a slight reduction in the intensity of the received signal.

7. Reduction of Interference in Broadcast Reception : A. N. GOLDSMITH : *Proc. Inst. Radio Eng.*, 1926, **14**, 575-603.

The local interference produced by a powerful telephony transmitter is discussed. Details are given of an investigation made upon a particular

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interfering station and it is shown that in most cases a simple form of wave-trap, or rejector circuit, will eliminate such local interference.

The paper is entirely of a pratical nature and no generalised work upon the subject has been undertaken.

8. **Radio Interference :** R. B. ASHBROOK and R. W. WRIGHT : *Elect. Wld.*, 1926, **88**, 851-53.

If adequate screening is adopted, no interference should be obtained when a radio receiver is situated directly beneath transmission lines up to 220 kilo-volts, providing the voltage on the line does not greatly exceed the corona point. The causes of interference are discussed and a receiver is described in which undesired frequencies due to D.C. generators are eliminated by means of a rejector circuit.

Several forms of rejector circuit are examined briefly. No generalised work is attempted and the paper is non-mathematical.

9. **The Performance of Reflexed Valves:** D. KINGSBURY : *Experimental Wireless*, 1927, **4**, 547-51.

The author describes a phenomenon which he observed with reflexed circuits, consisting of a howl produced due to a special form of interference between the radio-frequency and the audio-frequency impulses on the first grid.

(See also Abs. 1.1.1/7.)

10. **New Method pertaining to the Reduction of Interference in the Reception of Wireless Telegraphy and Telephony :** H. DE BELLESCIZE : *Proc. Inst. Radio Eng.*, 1926, **14**, 249-62.

The usefulness of directional reception and of resonant circuit selectivity in reducing the atmospheric effect of disturbances of reception is considered. The conditions under which weak sinusoidal E.M.F.'s will be completely suppressed, or ultimately delivered to the output, in the presence of much larger impulses are fully examined, and an analogy is drawn between the electrical cases and mechanical frictional and magnetic hysteretic systems.

The analysis is not very general. Circuit arrangements are suggested to eliminate interference.

11. **Resonance and Atmospherics :** H. DE BELLESCIZE : *Onde Élect.* 1927, **6**, 333-56, 427-44.

A lengthy discussion of resonance in receiving circuits and of the effect on such circuits of atmospherics leads the author to conclude that the action of atmospherics on receivers can be explained as due to disturbances which are more or less continuous, of fortuitous occurrence, and, at small or moderate distances, equally effective whatever the wave-length to which the receiver is tuned. Hence, an increase of the amplification or of the sensitivity of the receiver simply leads to an equal relative increase of the atmospherics.

12. **Broadcast Disturbance by Tramways and their Elimination :** A. CLAUSING and P. MÜLLER : *Elektrotech. Z.*, 1928, **49**, 178-80.

The methods of reducing the radio-frequency disturbances produced by the spark formation at the trolly wire are discussed. The methods are: (*a*) Using condensers between the trolly wire and earth ; (*b*) inserting radio-frequency chokes in the trolly wire ; (*c*) connecting condensers across the car lighting plant ; (*d*) increasing the pressure between and broadening the contact area of the trolly wire and collector ; and (*e*) choosing the most favourable material for the current collector. Oscillographic records show the dependence of the disturbances on the magnitude of the current and the collector material. Carbon collector pieces give the smallest disturbance.

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13. A Note on Some Interfering Oscillations Experienced in a Supersonic-Heterodyne Receiver : R. L. SMITH-ROSE : *Experimental Wireless*, 1928, **5**, 673-76.

These interferences were encountered in the development of a highly sensitive receiver for the reception of C.W. signals ; their origin and the methods of eliminating them are described. The intermittent action of a valve oscillator due to " squegging " causes are found particularly liable to occur if one valve is used for the dual purpose of the first oscillator and the first rectifier—a bad practice adopted by some manufacturers. A second kind is produced by the self-oscillation of the intermediate-frequency amplifier. A third, more subtle, kind can only be distinguished from an incoming signal by its lack of direction. It is due to the second local oscillator giving an impure oscillation, so that under certain conditions a harmonic from it gets into the primary or secondary receiving circuits.

14. Transmission and Reception by Super-Regenerative Receivers: G. BEAUVAIS : *Onde Élect.*, 1928, **7**, 206-9.

The author points out that two super-regenerative receivers working near to one another are mutually interfering, and that this feature may be of use, since it enables one station to communicate with the other by using its receiver as a low-power transmitter.

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1.3. RADIO-FREQUENCY TRANSFORMERS. ANALYSIS AND DESIGN

TRANSFORMER coupling between radio-frequency stages originally took the form of two closely coupled and tightly wound coils. Owing to the high effective resistance, large self-capacity, tight inductive coupling and large capacitative coupling, the selectivity as well as the amplification was very poor, particularly at very high radio frequencies.

The efficiency of medium-frequency transformers as used in intermediate-frequency supersonic-heterodyne amplifiers was improved by the practice of using thin laminated iron cores. If the iron is sufficiently thin to reduce eddy current losses, the selectivity can be quite good while the dimensions of the transformer are reduced by the use of iron. By suitably designing the iron path and the air gap, the response characteristic may be varied within wide limits, and if several stages are used, with slightly dissimilar tuning in each, a comparatively flat characteristic over a wide band of frequencies can be obtained. If the stages are all tuned to the same frequency, neutralisation must be adopted to obtain stability. The inductance of the transformer primary should be as large as possible, consistent with tuning requirements, together with a small step-up if possible.

For frequencies above 500 kilo-cycles per sec. (600 metres), highly efficient coupled tuned circuits are now adopted, but neutralisation must be resorted to, even for one stage, since the amplification is larger, due to such coils. With such transformers, in which the dimensions of the windings are necessarily considerable to obtain their efficiency, special precautions must be adopted to prevent inter-stage coupling due to stray magnetic fields. One method is to stagger the cylindrical coils and the other is to screen either the coil or the complete stage. The former method is not sufficiently effective for most multi-stage cases, the latter being preferable, though care must be exercised in designing the screen so that its proximity to the coil is never sufficient materially to increase the effective resistance of the coil due to losses in the screen.

A third method of eliminating stray effects is to use " toroid " transformers, in which the stray field is very small.

Considerable attention has been given to the design of the tuned radio-frequency transformer, particularly at high radio frequencies. The secondary of the transformer is shunted by the input impedance of the next stage, and this impedance materially modifies the performance of the transformer. It was not until the nature of this

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load was appreciated that the analysis of the behaviour of the transformer was put upon a sound basis.

For frequencies below 300 kilo-cycles per sec. (600 metres), air or iron-cored transformers may be used. If tight coupling is to be maintained, a double humped characteristic will be obtained and the circuit will be responsive to two frequencies. BLATTERMAN (see Abs. 1.3/4) has found that by suitably designing the iron circuit of an iron-cored transformer, almost any type of response characteristic can be obtained—a selective transformer with one frequency response maximum, or a transformer equally responsive to a limited band of frequencies and non-responsive to frequencies outside this band. This latter type is suitable for telephony reception where side bands need preservation. Such iron cored transformers are eminently suitable for the intermediate-frequency multi-stage amplifier of a supersonic-heterodyne receiver.

For frequencies above 500 kilo-cycles per sec. (600 metres), BUTTERWORTH'S work (see Abs. 1.3/6) on the design of coils has produced considerable improvements in efficient transformer design.

In modern receivers, the requirements are more of the nature of a filter circuit than a simple resonant circuit, i.e., a coupling arrangement which gives a constant amplification over a limited band of frequencies sufficient to cover the side bands accompanying telephony transmission. This requirement has been partially accomplished by the use of iron-cored transformers at frequencies below 500 kilo-cycles per sec. (600 metres), but no effective solution has been obtained for a small number of stages at higher frequencies. The effect of the load of the following stage is so serious that a design applicable to very high frequencies presents serious difficulties.

The requirements of selectivity and undistorted amplification are, unfortunately, not coincident, and it is doubtful whether, in view of the complicated nature of the input impedance of the following stages, the theoretical analysis can be entirely divorced from experimental investigation.

Satisfactory experimental results at frequencies above 1,000 kilo-cycles per sec. (300 metres) have not been obtained owing to measurement difficulties, and attention must be given to the design of measuring apparatus to obtain serious progress in design at these radio frequencies.

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1. **On Electrostatic Transformers and Coupling Coefficients:** F. C. BLAKE : *J. Amer. Inst. Elect. Eng.*, 1921, **40**, 23-29.

The problem of the capacitative coupling of two oscillating circuits is dealt with in this paper, utilising Rayleigh's equations and the idea of "elastance," which is defined as the inverse of capacity and is measured by the reciprocal of the farad, the "daraf." The effective elastance and resistance of a circuit is demonstrated to be a function not only of its own elastance, but also of the elastance, mutual elastance and self and mutual inductance of neighbouring circuits. In particular, beyond a certain frequency, the

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effective elastance, instead of increasing with increasing frequency, will, in general, decrease owing to the ratio of the reactance to the resistance steadily increasing in value, the effective elastance ultimately reaching the minimum given by the disappearance of the dissipative terms. The verification of the formulæ was accomplished by working with Lecher systems. The author explains, by the theory of elastically coupled circuits, the regenerative action of the triode, due to the elastic coupling between grid and anode, anode and filament, and grid and filament.

2. **Reaction of Rectifying Device on Transformers :** G. Joos : *Jahrb. drathl. Telegr.*, 1922, **19,** 276-80.

This paper gives a simple mathematical discussion of the effect of the valve input load introduced in the transformer secondary circuit on the primary circuit of a transformer.

3. **Radio-Frequency Amplifiers :** P. D. Lowell : *Sci. Pap. Bur. Stand. Wash.*, 1922, **18,** No. 449, 335-43.

After a general discussion of the relative merits of radio and audio-frequency amplifiers with various couplings between the stages, the author describes two types of transformer-coupled radio-frequency amplifiers. In the first, an extended frequency range is obtained by altering the coupling between the two transformer coils, and a device for doing this simultaneously on all stages is described. In the second, iron-cored transformers are used with stampings of silicon steel less than 0·001-inch thick.

An amplification characteristic plotted against wave-length is given but no attempt is made to conceal the fact that it is obtained by an approximate theoretical analysis only, as it is entirely unsupported by experimental verification and the scaled ordinates are very misleading.

4. **Notes on the Design of Radio-Frequency Intervalve Transformers using Iron Cores :** A. S. Blatterman : *Wireless Wld.*, 1922, **10,** 48-51, 78-84, 104-7.

With a pair of coupled circuits there are in general two frequencies at which maximum voltage is obtained across the terminals of the condenser in the secondary circuit. The separation between the peaks depends on several factors, of which the most important are the co-efficient of coupling and the internal resistance of the input valve. By properly proportioning the time-constants of the two circuits as well as the coefficient of coupling, the two peaks may be made to coalesce, or one may be relegated to zero frequency. In either case the system behaves as one responsive only to a single or double hump characteristic, depending upon whether or not the amplifier is required to be effective over a narrow or a relatively wide range of wave-lengths. The pre-determination of the coefficient of coupling is, therefore, of great importance. This involves the calculation of the leakage flux in the transformer and of the inductances of the transformer winding. It is found that practically any desired characteristic can be obtained by suitably adjusting the turns, the spacing between primary and secondary, and the air-gap in the core of the transformer. Formulæ are developed for the leakage flux, the coupling coefficient and the primary and secondary inductances in terms of the various dimensions and data concerned. The circuit problem is not dealt with, and it is shown how the free frequencies may be determined using a graphical method. Finally, an example of design is worked out numerically.

5. **An Efficient Tuned Radio-Frequency Transformer :** F. H. Drake and G. H. Browning : *Proc. Inst. Radio Eng.*, 1925, **13,** 767-79.

General formulæ are furnished for the design of a tuned radio-frequency transformer for any wave-length band, with special reference to the broadcast

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range. The secondary inductance, L_2, must be made as large as possible consistent with the ability to tune to the minimum wave-length desired. Capacitative coupling between the primary and secondary must be reduced to a negligible amount compared with the magnetic coupling K; by making K large, the region over which the amplifier tends to oscillate is reduced to a very narrow band adjacent to the setting for maximum voltage amplification. The theoretical values of voltage amplification are checked by experimental observation, the agreement being very good. If λ_0 (metres) is the minimum wave-length to which it is desired to tune, L_2 (millihenries) $= 5 \times 10^{-6} \lambda_0^2$. The primary inductance L_1 can then be determined, knowing also the anode resistance of the amplifier valve.

6. Effective Resistance of Inductance Coils at Radio-Frequency : S. BUTTERWORTH : *Experimental Wireless*, 1926, **3**, 203-10, 309-16, 417-24.

Formulæ are developed for the determination of the effective resistance of inductance coils at radio-frequencies and the best design of coils considered. Following the discussion of the alternating current resistances of straight wires, the author investigates the eddy-current losses in a cylinder placed in a uniform alternating magnetic field. These losses vary as the fourth power of the diameter and inversely as the resistivity at low radio frequencies, and as the first power of the diameter and inversely as the square root of the resistivity at high radio-frequencies. The author extends these formulæ to include single-layer coils for both high and low radio-frequencies and derives a formula which gives the best gauge of wire for any single-layer coil. The copper loss is usually the predominant factor, and such losses only are considered in this paper.

Multi-layer coils of small and finite winding depth are considered, and tables and curves are given constituting the most recent attempt at solution of the alternating current resistance problem for solid wire coils with well-spaced windings. The basis of design must be that for minimum copper loss and not from the point of view of dielectric loss, and the general principles for such design are given. The best coil for a given volume of copper is a single-layer coil having a winding depth equal to one-third of its diameter; designs for fixed length of wire and fixed overall surface are also included.

This paper is inserted in this section because the design of radio-frequency coils for transformers has been materially advanced by the findings of this author.

7. Supersonic Transformers : N. W. MCLACHLAN : *Wireless Wld.*, 1926, **19**, 631-4, 680-4, 715-8, 777-8; *Wireless Wld.*, 1927, **20**, 21-24.

The author gives some results obtained with various designs of thin iron-cored transformers at wave-lengths of 5,000 to 28,000 metres, and compares them with the results of an American transformer with practically no iron. He states that :—

(1) Iron is permissible in intervalve transformers for currents of supersonic frequencies.

(2) At any given optimum wave-length, the breadth of the top of the amplification-frequency curve is greater, the lower the internal resistance of the valve.

(3) The breadth of the top increases with decrease in the optimum wave-length.

(4) The effective primary inductance should be large and the capacity small.

(5) The thinner the magnetic material and the higher the quality of the iron, the shorter the optimum wave-length.

(6) With large leakage (50 per cent.) and a valve of moderate internal resistance, the amplification curve is flattened.

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The effect of the input impedance of valves on the performance of 1·5 mil. iron transformers is discussed. A resistance-coupled stage following a transformer considerably augments the operating wave-length, and staggered windings are almost useless unless the valve capacity is neutralised. In condenser-resistance amplifiers, inter-electrode capacities reduce the amplification, especially at very high frequencies.

The author also points out that isolated curves of one unloaded transformer are of little value where another transformer and a selective unit are also used, the overall amplification of the complete stage being required.

8. **Improvements in Tuned Amplifiers :** BLANCHARD : *Onde Élect.*, 1927, **6**, 57-70.

Two circuit arrangements are described in which the amplification is practically constant over a small range of frequencies and falls off very rapidly outside this range.

These arrangements, although not filters, have similar properties and are very useful as intermediate-frequency transformers for supersonic-heterodyne receivers.

9. **Radio-Frequency Transformers :** N. W. MCLACHLAN : *Experimental Wireless*, 1927, **4**, 597-600.

The article is an approximate mathematical analysis of the radio-frequency transformer, and shows that for a given valve there are certain optimum conditions of coupling and ratio of turns. The application of this analysis to the case of the screen-grid valve shows that in this case the best transformer only gives a gain of 8 per cent. in magnification over the tuned anode type of coupling.

The paper is purely theoretical and no experimental results are given in confirmation of the theory.

10. **A Mathematical Study of Radio-Frequency Amplification :** V. G. SMITH : *Proc. Inst. Radio Eng.*, 1927, **15**, 525-36.

The author investigates the fundamental circuit of the tuned transformer. If $R_1\ X_1\ Z_1$; $R_2\ X_2\ Z_2$ denote the resistance, reactance and impedance of the primary and secondary circuits respectively and M the mutual inductance, the conditions for maximum secondary current are

$$\frac{X_1}{R_1} = \frac{X_2}{R_2} \text{ and } M^2\omega^2; = Z_1Z_2$$

which may be established with only two variables, making it unnecessary to tune the primary. The secondary current is then

$$\frac{E_1}{2\sqrt{R_1R_2}}$$

This leads to a valve voltage factor of

$$\frac{\mu}{\sqrt{R_2}}$$

as a measure of its efficiency, and a coil constant $\frac{\omega L_2}{\sqrt{R_2}}$ for the secondary coil efficiency. When only the secondary is tuned, the maximum secondary current occurs when

$$X_2 = \frac{M^2\omega^2}{Z_1^2} X_1 .$$

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Curves representing this current are given and discussed. The effect of the valve output capacity is the same as though the valve had a complex amplification factor and a complex anode impedance.

11. **Selectivity of Tuned Radio Receiver Sets :** K. JARVIS : *Proc. Inst. Radio Eng.*, 1927, **15**, 401-23.

Decreasing the transformer primary turns increases the selectivity, but the maximum selectivity thus obtainable is only double that at optimum amplification. If the secondary inductance is adjusted for optimum amplification, the circuit is also adjusted for optimum selectivity ; increasing the secondary inductance increases the selectivity. Several curves illustrating these points are given. The factors affecting the quality are discussed, in particular the non-uniform amplification of the side bands.

12. **Inductive Coupling with Primary and Secondary Tuning Connecting Valves :** H. KAFKA : *Jahrb. drahtl. Telegr.*, 1927, **30**, 44-52.

The effect of the inductive coupling in a two valve circuit is theoretically investigated. The internal resistance of the first valve exerts a considerable influence on the effect of the coupling. With secondary tuning, the sharpness of resonance and the magnitude of the grid voltage are affected by the degree of coupling.

The influence of the coupling on the magnitude of the grid voltage of the second valve is represented by a locus diagram, and the voltage attains a maximum value for a definite degree of coupling. Finally, the author investigates how far loose coupling can be practically employed to obtain an increased sharpness of tuning.

The experiments described are intended to be a starting point for the design of inductive couplings to connect valves in the case of radio and intermediate-frequency amplifiers.

13. **Tuned Radio-Frequency Amplifiers :** R. S. GLASGOW : *J. Amer. Inst. Elect. Eng.*, 1928, **47**, 327-31.

(See Abs. 1.2.2/20.)

14. **Data on the Voltage Amplification of Radio-Frequency Transformers :** B. K. OSBORN : *Radio Eng.*, 1928, **8**, 13-15.

An analysis of the measured performance of several transformers is given. The transformers tested are classed as " continuous winding " if the direction of rotation is the same as one proceeds from the anode of the amplifier valve, through the primary, to the filament end of the secondary and through the secondary to the grid of the output valve ; and as " reversed " if the direction of rotation changes as one passes from primary to secondary. Curves are given showing that the " reversed " winding gives the higher amplification at the maximum point on the curves, moreover, it gives greater selectivity. These effects are not produced in transformers so wound as to have very low capacity between the two windings. Other effects of variation in the methods of winding, etc., are shown.

15. **Note on Radio-Frequency Transformer Theory :** H. DIAMOND and E. Z. STOWELL : *Proc. Inst. Radio Eng.*, 1928, **16**, 1194-1202.

In the mathematical analysis of radio-frequency amplifiers employing tuned-transformer coupling, the effect of the distributed capacitative coupling existing between the transformer windings has heretofore been neglected. General equations including this effect are developed in this paper. It is shown that with these equations, a closer agreement between computed and experimental results can be obtained. The presence of capacitative coupling changes the amount of mutual inductance necessary to obtain optimum

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amplification at a given frequency, and at the same time reduces the optimum voltage gain. The manner of variation with frequency of the equivalent transformer primary impedance is also altered.

16. **Aperiodic H.F. Amplification with Modern Valves :** F. L. DEVEREUX : *Wireless Wld.*, 1929, **24**, 173-4.

The author offers suggestions for the application of recent developments in valve and radio-frequency choke design. Attempts at radio-frequency amplification with radio-frequency choke coupling have hitherto met with little success for wave-lengths much less than 1,000 metres. The advent of modern valves, with a rated amplification factor of 15 and an A.C. resistance of as low as 7,500 ohms, and the general improvements in the characteristics of commercial radio-frequency chokes, make it worth while to reconsider this method of amplification, particularly as a preliminary stage in super-heterodyne or portable receivers.

It has been found possible to design a single stage with choke capacity-coupling which has a sensibly uniform characteristic over the frequency range of 100 to 1,000 kilo-cycles per sec. (300 to 3,000 metres), but it appears that two stages do not give such desirable characteristics owing to the approach of resonance in the R.F. chokes.

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1.4. RETRO-ACTION AND STABILISATION

THE complicated effects due to inductive and capacitative coupling between various parts of an amplifier present great analytical difficulty. The theoretical examination of these effects has been confined entirely to special coupling arrangements and to the effect of the anode-grid capacity.

Retro-active effects are due to the deliberate retro-active device (if one is inserted) and to the unavoidable inherent retro-action that takes place in the amplifier circuits.

Deliberate retro-action devices may take the form of—

(*a*) Inductive coupling coils.
(*b*) Condensers connected between anodes and grids of certain valves in a multi-stage case.
(*c*) Condensers in conjunction with tapped inductances giving critical retro-action control to one stage. (Reinartz Circuit.)

Inductive coupling devices, which feed a small e.m.f. derived from the output into the input circuit with such a phase relationship that the input is stimulated, are the commonest form of retro-action device. Critical adjustment is usually difficult and the re-tuning of the primary circuit made necessary by the alteration of the mutual inductance between the primary and the retro-action coil introduces a serious practical objection. Capacitative retro-action is preferable and may be utilised without the above-mentioned disadvantages. Small condensers, coupling anodes to grids, are often used to stimulate self-oscillation, care being taken that the phasing of such devices is correct. Perhaps the best of these capacitative retro-action controls is the Reinartz system, where critical retro-action is obtained without any serious change in the primary tuning.

The object of all such deliberate retro-active devices is to reduce the input circuit resistance, improve the selectivity and obtain a greatly improved performance from the receiver. The effects of too much retro-action are well known, namely, the cutting off of side band frequencies due to a resonance curve which is too peaky.

A general mathematical analysis of such a retro-active scheme, in which the amplifier is assumed to be perfect and all retro-action to be deliberately inserted externally as a unit, has been given by BRILLOUIN, GUTTON, FILIPPINI and THOMAS (see Abs. 1.4.1/2, 3, 12 and 19). A graphical treatment is given and the conditions for stability and retro-action are obtained, the analysis only serving as a guide to an understanding of general properties. The assumptions made are of course too general to apply to a particular case.

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The theory of such retro-active devices is well known, and, provided quantitative precision is not demanded, the existing theory explains the experimental evidence satisfactorily.

The unavoidable residual retro-action due to the couplings within the amplifier, which render it unstable under certain conditions, may be classified as follows (1.4.2) :—

(*a*) Inductive coupling between the output and the input circuits.
(*b*) Capacitative coupling between the output and the input circuits.
(*c*) Capacitative coupling between individual stages.
(*d*) Inductive coupling between individual stages.
(*e*) Coupling between stages due to battery resistance.
(*f*) Inherent valve inter-electrode coupling

(*a*) and (*b*) can be reduced by reasonable care in design and (*e*) and (*d*) can be reduced by suitable anode-feed systems and " toroid " or screened coils respectively. It is difficult to reduce (*c*) below a certain limit imposed by the wiring, valve holders, etc., though stage screening is beneficial, but it is impossible to eliminate (*f*) if a triode is used, and herein lies the difficulty of analysing the performance of an amplifier, since the inter-electrode capacities not only produce a shunt load on the input system, but also give rise to regenerative effects. The theoretical and experimental work on this subject is scanty and of an elementary character only, the application of general analytical methods being very difficult.

The instability due to these six causes has received much practical consideration and many devices have been produced to neutralise these retro-active effects. The earliest of these devices consisted in deliberately inserted grid damping by giving the grid a slight positive potential, and another similar method consisted in winding the coupling transformers with resistance wire or inserting a variable damping resistance. The difficulty in attempts to produce stability by damping methods of this kind lies in the fact that a valve may generate oscillations although its amplification is small, and the circuits so produced are inferior as to selectivity and amplification without materially reducing the liability of self-oscillation.

To prevent magnetic coupling between coils, staggering, the use of " toroids " and screening either coils or stages may be resorted to, but such screens, if too near to the windings, may increase the radio-frequency resistance of the coils with consequent loss of amplification and selectivity.

Perhaps the most successful method is that known as " Neutralisation " in which the anode-grid inter-electrode capacity is balanced by means of a similar small external condenser forming part of a bridge network. By this means stability can be obtained, but since the anode-grid impedance is not a pure reactance owing to the circuit loading, its value will depend upon the frequency, and consequently no perfect balance can be obtained with a pure capacity

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as balancing agent. Attempts have been made to overcome this difficulty by shunting the neutralising condenser with a high resistance to imitate the anode-grid impedance, but a better method is a balanced push-pull arrangement employing two valves of identical type. In practice, however, this arrangement becomes difficult unless the valves and circuits are identical.

The complete theory of such neutralised circuits has not been developed and the experimental results obtained with a neutralised amplifier show that there are big discrepancies between the measured results and any simple neutrodyne theory.

At very high radio-frequencies, neutralisation has been used, but here, no experimental work of a precise nature is forthcoming.

It appears probable that the action of stabilising devices cannot be explained in terms of any simple theory and there is need of a thorough investigation of this subject, particularly at very high radio-frequencies.

Undoubtedly the most promising method of obtaining stability is by the use of the screen-grid valve. The problem is here attacked at its source, viz., instead of attempting to neutralise the effects of the inter-electrode capacities, the inter-electrode capacity is very materially reduced by internal screening. Such methods are giving excellent results, but neutralisation cannot entirely be dispensed with. Although it is possible to obtain stability with an un-neutralised screen-grid stage, it is found that a greatly improved performance can be obtained if the last trace of inter-electrode capacity is balanced out by a neutralising scheme. The problem is therefore one of degree only, and the theory of neutralisation must be applied to the screen-grid valve in the same manner as it is applied to the triode stage.

BIBLIOGRAPHY

1.4.1. General Theory

1. Characteristic Curves and the Conditions for Stable Operation : J. Béthenod : *Lumière Élect.*, 1916, **35**, 25-31.

The paper discusses the elementary conditions necessary for a valve network to remain stable.

The discussion is of a very general mathematical nature and is, in fact, applied to mechanical systems. It is not readily applicable to the solution of practical cases owing to the neglect of many factors which would seriously affect the results.

2. Resistance Amplifiers : L. Brillouin : *Onde Élect.*, 1922, **1**, 7-17, 101-23.

The essential features of the resistance-capacity-coupled amplifier are stated and a very simple explanation of rectification is given which, however, excludes the loading effect of the detector on the preceding circuit. An experimental investigation of various problems connected with the construction of multi-stage amplifiers is reported, particular attention being given to capacity " compensation " coupling between the stages. The

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stability of the system is obtained by a suitable small capacitative coupling between the grid of the first valve and any particular anode in the multi-stage amplifier. A theoretical examination is given of the phases existing between the E.M.F. fed back to the grid and the original input, and the condition for self-oscillation is calculated. The simple analysis of internal resonance is given and the nature of the retro-action effect produced by the anode-grid inter-electrode capacity is examined. The design of a practical six-stage amplifier with capacity stabilisation is given.

3. **On the Building Up and Sustaining of Oscillations in an Amplifier :** C. GUTTON : *Onde Élect.*, 1922, **1**, 261-70.

The vector method has here been extended to any form of coupling between the output and input circuits, and the simple theory of inductive coupling between these circuits is given.

The action of the retro-active coupling in a triode circuit for reinforcing, damping, or generating oscillations is examined by considering the triode as a relay. The conditions of working in each case are drawn out and shown to be identical with those directly obtained from the differential equations of the triode.

4. **Radio Receiving Equipment :** F. CONRAD : *Proc. Inst. Radio Eng.*, 1922, **10**, 426-39.

Part of this paper shows that the increase of signal strength which can be obtained by the use of retro-action is limited to a definite value, which is determined by the strength of the incoming field, and it is shown by curves that this maximum signal strength is independent of antenna height.

The paper is non-mathematical and no generalised theory is attempted.

(See also Abs. 1.2.10/3.)

5. **Resistance Neutralisation :** E. BENNETT and L. PETERS : *J. Amer. Inst. Elect. Eng.*, 1922, **41**, 234-48.

The paper deals in a very general manner with the effect of retro-action, and it is shown that the retro-acting device not only supplies power to the circuit but causes the source of driving E.M.F. to furnish more power to the circuit than it would if the retro-acting device were not present. The improved selectivity of the system is shown to depend upon the intensity of the regenerative effect. The manner in which a triode may function as a retro-acting device is examined and general examples of a very theoretical nature are given.

The theory is of a very elementary nature and no experimental data is available.

6. **Retro-action in Tuned Amplifiers :** H. DE BELLESCIZE : *Radio électricité*, 1922, **3**, 183-7, 238-42.

In amplifiers containing several triodes and several oscillating circuits, special phenomena often take place, the explanation of which is not always simple. Sometimes the tunings of the different circuits are dependent upon one another, sometimes some of the circuits appear to have acquired an extraordinarily sharp resonance, and in other cases the intensity of reception reaches a maximum for two different values of the capacity of a condenser.

In the present article the theory of such tuned amplifiers is given and an attempt is made to explain the phenomena cited above.

7. **High Frequency Reaction Amplifiers :** L. DUBAR : *Onde Élect.*, 1923, **2**, 43-49.

In the first section, an elementary explanation is given of an irreversible lagging effect of oscillations encountered in radio-frequency amplifiers.

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The second part gives a short theoretical treatment of retro-action applied to radio-frequency amplifiers, either transformer coupled or with resistance-capacity coupling between the valves.

In the third section, a brief account is given of an autodyne amplifier with four valves and with capacitative retro-action between the anode of the third valve and the grid of the second.

It is stated that this arrangement gives as good a result over the wave-length range, 2,000 to 5,000 metres, as can be obtained when using a separate heterodyne.

8. **Quantitative Study of Inductive Reaction :** C. B. JOLLIFFE and J. A. RODMAN : *Sci. Pap. Bur. Stand. Wash.*, 1924, No. 487, 419-28.

Simple alternating current theory is used to show that the amplification produced by the type of retro-action employing inductive coupling between the anode and tuned grid circuits of a receiver, can be considered as caused by a reduction in the resistance of the tuned circuit. An expression is obtained for the current amplification with inductive retro-action, confirmed by experiments in which the current in the tuned circuit with and without retro-action was measured by means of a valve voltmeter which measured the drop of potential across the condenser in the tuned circuit.

It is stated that the presence of the valve voltmeter did not affect the tuned system, though no details are given of the conditions under which the voltmeter was used. Serious errors can be introduced by such an arrangement unless special precautions are taken.

9. **Back-Coupling in a Four-Valve High-Frequency Amplifier :** W. HEY : *Jahrb. drahtl. Telegr.*, 1924, **23**, 56.

The amplifier here described has a tuned circuit in the anode of the first valve, with capacitative coupling to the grid of the second valve. Choke-capacity coupling is used between the second and third, and between the third and fourth valve. The choke in the anode lead of the third valve is variably coupled to the inductance of the tuned anode circuit of the first valve. In this way self-oscillation can be brought about and continuous-wave stations can be received. Back-coupling the anode choke of the third valve to part of the inductance of the receiving circuit gives similar results. It is stated that with the above arrangement signals having a frequency of 500 to 12·5 kilo-cycles per sec. (600 to 24,000 metres) have been received.

10. **The Limit of Regeneration :** N. C. LITTLE : *Proc. Inst. Radio Eng.*, 1924, **12**, 479-83.

The maximum retro-action possible before oscillation is produced in a simple circuit with a retro-acting coil in the anode circuit of the valve was obtained. Two retro-acting coils were used, one with a large number of turns and one with one turn only for fine adjustment. It is suggested that this arrangement should be adopted in receivers using critical retro-action to prevent over-stepping the oscillation point due to the coarseness of the retro-active control.

The paper is non-mathematical and merely gives practical details regarding the construction of inductive retro-acting controls.

11. **Regeneration in Coupled Circuits :** E. L. CHAFFEE : *Proc. Inst. Radio Eng.*, 1924, **12**, 299-359.

In Part I, the simple theory of retro-action in a circuit is developed (1) when the oscillatory circuit is included in the anode lead and (2) when it is in the grid circuit.

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In Part II, the theory of coupled circuits having positive resistances is dealt with and is illustrated by means of space models for the secondary current.

In Part III, the theory of retro-action in coupled circuits is developed. It is shown that retro-action in the secondary circuit only affects the impedance of that circuit, and may make the equivalent resistance of that circuit zero. Expressions are obtained giving the relative selectivity of coupled circuits, and curves are given showing the results in two numerical examples.

12. **Contribution to the Theory of Amplification and Reaction :** G. FILIPPINI : *Elettrotechnica*, 1924, **11**, 914-22.

The analysis of GUTTON (see Abs. 1.4.1/3) is here considerably extended and graphical circle diagram methods are used to show the optimum conditions of coupling.

The circle diagram is made use of and a more complete examination of the inductively coupled case is given.

13. **Some Thermionic Valve Problems :** E. V. APPLETON : *Rep. Brit. Ass.*, 1925 ; *Electrician*, 1925, **95**, 266-7.

(See Abs. 1.2.1/17.)

14. **The Effect of Stray Reactions on the Stability and Amplifying Power of Amplifiers :** O. F. BROWN : *Experimental Wireless*, 1925, **2**, 434-40.

A review of causes of instability in multi-stage amplifiers.

Damping is one of the methods adopted to prevent self-oscillation, either by (1) increase of grid potential, which involves increasing the grid current and introduces a finite resistance across the grid circuit, and (2) the introduction of resistance into the circuit or winding the intervalve transformers with resistance wire. The difficulty in attempts of this kind lies in the fact that a valve may generate oscillations although its amplification is small, and the circuits so produced are inferior as to selectivity and amplification, without materially reducing the liability of self-excitation. Reduction of undesired couplings between the circuits may be secured by enclosing the coils in metal screens, winding the coils as practically closed rings (toroids), or winding the coils on cylinders which are staggered with respect to each other. Various methods of overcoming the difficulty are described in the paper.

Little actual experimental data is available.

15. **An Analysis of Regenerative Amplification :** V. D. LANDON and K. W. JARVIS : *Proc. Inst. Radio Eng.*, 1925, **13**, 709-53.

An important paper on the principles and use of retro-action as an amplifying agency in a receiver.

After criticising the defects of the usual method of treatment, the authors develop an explanation on a basis of power balance. The analysis is first applied to the case of a tuned anode stage capacitatively-coupled, with retro-action from the anode of the rectifier to the tuned anode circuit, and the application of the same analysis to the case of retro-action to the aerial is indicated at the end of the paper. The treatment consists in considering the power losses involved without retro-action, and how these are modified when retro-action is applied. The reasoning is well illustrated by curves. It is shown that in the absence of a signal and with insufficient retro-action to render the system unstable, the sum of the losses considered is greater than the power supplied by the retro-acting coil. A signal voltage then prevents certain losses from occurring, thus upsetting the power balance

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between the power supplied by the retro-action coil and the power losses of the circuit. This causes self-oscillation to occur, so that effectively retro-action consists of self-oscillations started and controlled by the signal. The limit of amplification obtained in this way is due to variations in the anode and grid impedances, as the amplitude of the grid voltage increases. The effects of these variations in terms of power losses are shown in curves. The effects of a grid-leak and condenser are similarly dealt with in terms of power. It is concluded that they decrease the voltage amplification, but that increased rectification efficiency more than compensates for the difference when the audio-frequency output from the rectifier is considered. Resistance in the grid circuit decreases the amplification by increasing the effect of the impedance variations. The effect of retro-action upon distortion is considered and it is shown that a low L/C ratio may cause a loss of high audible notes if a weak signal is being received with full retro-action. It is shown that amplification due to the effect of retro-action still occurs even when the valve is in a condition of weak self-oscillation, this applying both to audio beating and zero beating. Zero beating is illustrated by a series of vector diagrams. Lastly, the sharpening of tuning due to retro-action is considered.

16. **Performance of Amplifiers :** H. A. THOMAS : *J. Instn. Elect. Eng.*, 1926, **64**, 253-68, Disc. 268-78 ; *Experimental Wireless*, 1926, **3**, 15-18 ; *Elect. Rev. Lond.*, 1926, **98**, 73-74.

The section of this paper applicable under this heading gives a theoretical examination of the effect of inductive retro-action, and experimental results show the reduction in effective resistance of the input circuit due to this form of retro-action.

(See also Abs. 1.2.7/6, 3.4/9, 4.2/8, 4.4.1/7, 4.5/7, 4.6/1.)

17. **The Alteration of Resistance, Inductance and Capacity by Means of a Resistance Coupled Amplifier :** C. AIKEN : *J. Opt. Soc. Amer.*, 1927, **15**, 85-95.

A simple mathematical treatment of retro-action.

18. **Reaction and Selectivity :** A. P. CASTELLAIN : *Wireless Wld.*, 1927, **21**, 757-8.

This paper points out the advantages obtained by the use of efficient coils in preference to the use of retro-action to reduce the input circuit resistance.

The analysis is very elementary and the experimental results are open to grave criticism on the score of loading due to the use of a valve voltmeter connected across the tuning condenser.

19. **Retro-action in Amplifiers :** H. A. THOMAS : *Experimental Wireless*, 1928, **5**, 245-51.

Retro-action in an amplifier is the property of feeding back energy from the output to the input, and the consequent modification in the normal amplification which it would possess if no such feed-back took place. The author examines the effect mathematically in terms of a retro-action coefficient, and shows how to determine by vectorial methods the amplification obtained.

The paper represents an extension of the vector method of solution suggested by GUTTON and FILIPPINI.

(See Abs. 1.4.1/3, 1.4.1/12.)

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20. **The Stability of a Valve Amplifier with Tuned Circuits and Internal Reaction :** R. T. BEATTY : *Proc. Phys. Soc. Lond.*, 1928, **40**, 261-8.

A mathematical paper dealing with the stability of amplifiers under several conditions. The case considered is that in which both the input and output circuits are tuned and the inter-electrode capacity is the only source of retro-action. The final result is in complete agreement with that of POSTHUMUS for the special case of a repeating amplifier, but differs in phase relations from the result of HULL. Two examples are given to show how the result, given in graphical form, may be applied to specific cases. It is shown also that an increase in the grid-anode conductance of the valve leads to greater stability, with a decrease in amplification.

21. **Reaction Amplifiers :** Y. WATANABE : *Jahrb. drahtl. Telegr.*, 1928, **32**, 77-83.

Equations are theoretically developed which show the relation between amplification factor, limits of amplification without oscillation and degree of retro-action for all types of retro-active coupling.

22. **Resistance Back Coupling in Valve Amplifiers :** F. SCHIERL : *Arch. Elektrotech.*, 1928, **20**, 346-70.

(See Abs. 1.2.8/13.)

23. **The Effect of Regeneration on the Received Signal Strength :** B. VAN DER POL : *Proc. Inst. Radio Eng.*, 1928, **16**, 1045-52.

The author gives a theory of the effect of retro-action, using the solution of a non-linear differential equation, and gives an experimental verification of the theory. It is shown that (*a*) as a first approximation, rectification has no effect on the radio-frequency grid voltage developed under the influence of an incoming signal ; (*b*) the amplification obtained through the use of retro-action equals the two-thirds power of the ratio of the " grid-space " to the amplitude obtained with zero retro-action. It is apparent from (*b*) that much greater gain is obtained by retro-action with weak signals than with strong signals. The verification of the theory is made with a circuit arrangement operating at 500 cycles per second. Application is made to radio-frequencies using this as a model of a radio-frequency system, following a theorem for model systems which is stated.

24. **A Method of Treating Resistance-Stabilised Radio-Frequency Amplifying Circuits :** B. L. SNAVELY and J. S. WEBB : *Proc. Inst. Radio Eng.*, 1929, **17**, 118-26.

A radio-frequency amplifying circuit may be stabilised by the insertion of a resistance in the grid circuit. If the value of this resistance exceeds the value of the negative input resistance of the valve, self-oscillation cannot occur. Since the input capacity of the valve is generally not negligible in comparison with the tuned circuit capacity, any increase in the grid-circuit resistance beyond the critical value causes an effective increase in the resistance of that circuit.

An equation is developed for a resistance-stabilised amplifier having a tuned grid circuit and any type of anode load. The critical value of this resistance is so dependent on the grid-filament capacity that a method of measuring small capacities is suggested, the unknown capacity being connected between the grid and filament. A variable condenser thus connected will provide a control of retro-action in such a circuit.

This suggested method of measuring small capacities has not been developed in the paper, which is entirely of a theoretical nature.

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1.4.2. Instability due to Inter-Electrode Capacity Effects and Theory of Neutralisation. (The "Neutrodyne")

1. **Regeneration and Oscillation in Vacuum Tube Circuits through Inter-Electrode Tube Capacity :** A. S. Blatterman : *Radio Rev.*, 1920, **1**, 633-43.

The purpose of this paper is to show the importance of the inter-electrode capacity of valves and to investigate what types of loads in the grid and anode circuits are required for self-oscillation to occur, the only coupling between these circuits being constituted by the inter-electrode capacities. The problem is dealt with mathematically and as a result it is found that regenerative feed-back can take place in four cases, namely : (1) when the anode circuit is purely resistive, whilst the grid circuit is purely inductive ; (2) when both circuits are purely inductive ; (3) when the anode circuit is purely inductive whilst the grid circuit is purely resistive ; and (4) when the grid circuit is purely inductive and the anode circuit is purely capacitative.

The conditions for maximum retro-action in the four above-cited cases are then mathematically determined.

2. **Resistance Amplifiers :** L. Brillouin : *Onde Élect.*, 1922, **1**, 7-17, 101-23.

(See Abs. 1.4.1/2.)

3. **Vacuum Tubes and their Technical Applications :** H. Barkhausen : *Jahrb. drahtl. Telegr.*, 1920, **16**, 82-111.

The effect of the anode-grid capacity as a means of producing retro-action is examined and the conditions for instability are theoretically stated.

There is no experimental confirmation.

(See also Abs. 1.2.2/7.)

4. **The Neutrodyne :** W. James : *Wireless Wld.*, 1923, **12**, 67-71.

The paper gives a simple discussion of the theory of the neutrodyne receiver, and details for the construction of such a receiver are appended.

5. **Self-Excitation of Valves through the Grid-Anode Capacity :** H. Barkhausen : *Jahrb. drahtl. Telegr.*, 1923, **21**, 198-221.

The paper gives mathematical and experimental work on the relations that must exist between the grid-anode capacity, the valve voltage factors, and the impedances of the external anode and grid circuits in order that self-oscillation may occur.

6. **Tuned Radio-Frequency Amplification with Neutralisation of Capacity Coupling :** L. A. Hazeltine : *Proc. Radio Club Amer.*, 1923, **2**, No. 8.

The manner in which the self-capacity of a coil may be neutralised is examined first, and then applied to the more general case of the balancing of two complex circuits. The application of this capacity neutralisation to an amplifier is then examined, and the paper concludes with a statement and a description of various forms of neutralisation, and a description of the commercial form of the receiver.

This is Hazeltine's original paper on the neutrodyne circuit.

7. **The Amplification and Selectivity of a Neutralised Tuned Anode Circuit :** N. W. McLachlan : *Experimental Wireless*, 1926, **3**, 545-552.

(See Abs. 1.2.2/18.)

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8. **Calculation of Apparent Capacity in Resistance Amplifiers :** M. v. ARDENNE and W. STOFF : *Jahrb. drahtl. Telegr.*, 1927, **30**, 86-89.

(See Abs. 1.2.7/8.)

9. **Stability of an Amplifying System :** K. POSTHUMUS : *Tijdschr. ned. Radiogenoot.*, 1927, **3**, 106-12.

The stability of a system consisting of n triodes is worked out, with attention to the inter-electrode capacities. The tendency to oscillation is hardly increased by raising the number of triodes above two or three.

10. **Detuned High Frequency Amplifiers :** W. RUNGE : *Telefunkenztg.*, 1927, **9**, No. 47, 50-63.

A theoretical discussion of the neutrodyne circuit.

11. **The Stability of the Tuned-Grid, Tuned-Plate High Frequency Amplifier :** R. T. BEATTY : *Experimental Wireless*, 1928, **5**, 3-15.

An expression is first derived giving the overall amplification produced by a single stage of amplification, and its geometrical interpretation is given ; the threshold of stability and voltage gain are discussed in some detail.

The paper is mathematical and cannot readily be applied to practical cases.

12. **Compensation of Capacities and their Reactions in Electron Tubes :** M. v. ARDENNE and W. STOFF : *Jahrb. drahtl. Telegr.*, 1928, **31**, 122-8, 152-7.

Four fundamental methods of eliminating the tendency to oscillate in radio-frequency amplifiers with tuned couplings are investigated. In the divided anode circuit of HAZELTINE an almost complete compensation of capacity retro-action is obtained. An improvement in this circuit is only possible by variation of the neutralising condenser and simultaneous detuning of the tuning circuit, but can be obtained with the divided grid circuit of LEITHAUSER without detuning. A combination of both fundamental circuits offers no advantage. The elimination of the tendency to oscillate for all frequencies is only possible by adopting a radio-frequency balanced push-pull arrangement.

13. **Harmful Effects of Inter-electrode Capacity :** M. v. ARDENNE and W. STOFF : *Experimental Wireless*, 1928, **5**, 509-13.

The theory of the effect of the inter-electrode capacity in introducing retro-action from the anode-circuit load to the grid circuit is given in some detail. It is pointed out that amplifier calculations can only be made by starting at the output valve and working backwards stage by stage. A pure resistance in the anode circuit causes an additional capacitative impedance in the grid circuit. Capacitative and inductive anode loads cause positive and negative resistance components respectively in the grid impedance.

The mathematical expression of the results is not new and the paper is essentially a concise statement of the simpler problems of input impedance.

14. **The Problem of the Most Economic Multiple Frequency changing in Reception :** F. AIGNER : *Jahrb. drahtl. Telegr.*, 1929, **33**, 9-15.

(See Abs. 1.2.4/23.)

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SECTION II—RECTIFIERS

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2.1. GENERAL THEORY OF RECTIFICATION

THIS section deals with the theory of rectification in its most general form, without reference to any particular type of rectifier. That is to say, it is concerned with the theory of an assumed conductor of electricity having a non-linear current-voltage characteristic. Papers which, although entirely theoretical in character, refer specifically to any actual type of rectifier will be considered only in their appropriate sub-sections.

Comparatively little work has been published on this subject. In fact only six papers of any importance have been found after a thorough search of the literature on rectification, and these taken together cannot be said to constitute more than an introduction to the subject.

One of the earliest attempts at the presentation of the theory of a non-linear characteristic is that by LIEBOWITZ in his paper on "Quantitative Relations in Detector Circuits" (see Abs. 2.1./1), where an analysis is given of a "perfect" rectifier (i.e., zero conductivity in one direction and finite and constant conductivity in the other), and also of a so-called "approximate" rectifier (i.e., one having very small conductivity in one direction and finite conductivity with a curved characteristic in the other). The paper is more particularly concerned with heterodyne reception, and the conclusions arrived at are of limited practical applicability owing to the omission of the important factor of the effect of the load on the behaviour of the rectifier. Within this limitation, however, the paper contains some useful analytical results of the application of Fourier analysis to the types of characteristic considered.

The first attempt at a completely general theory of the subject in which the effect of the output load is taken into account appears to be that given by COLEBROOK in his paper on "The Rectifying Detector" (see Abs. 2.1/3).

Here, the rectifying characteristic is specified as a power series (though consideration is also given to the limiting case of "perfect" rectification, i.e., zero conductivity in one direction, and finite and constant conductivity in the other), and the rectification performance is described in terms of an effective input resistance and

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an effective output rectified voltage associated with a certain effective internal resistance. The magnitude of these quantities, and their variation with operating conditions, is given in terms of the assumed power series coefficients of the characteristic. The analysis is not entirely satisfactory from a practical point of view, but this scheme of specification appears to be of some value from the point of view of design.

CHAFFEE and BROWNING (see Abs. 2.1/4) have given a fairly complete application of these ideas to continuous wave and modulated continuous wave rectification, but the utility of their presentation is limited to the case of signals so small that the characteristic can be considered as parabolic over the range of variation concerned. This is rather a severe limitation in practice.

A possible alternative form of systematisation, the full possibilities of which have yet to be investigated, is the specification of a rectifier in terms of so-called " dynamic characteristics," which are in effect lines recording the continuous wave rectification performance with no output circuit load. This is proposed by P. DAVID (see Abs. 2.1/5) as a sufficient source of data for the determination of the behaviour of the rectifier under assigned conditions of operation. In the absence of a more complete investigation, it cannot be stated whether such a specification is either sufficient or convenient, but the suggestion is worthy of consideration. (The term " dynamic characteristic " is not very suitable for the significance that DAVID attaches to it. The term is more appropriately applied to the current-voltage relationships of the detector when the currents and voltages are changing instantaneous values, these characteristics being thus distinguished from " static characteristics," representing the current-voltage relationships for steady current and voltage conditions. The term " dynamic characteristic" will be used in this sense throughout this section.)

The remaining papers are of less general importance, being concerned with the analysis of certain special cases. That by BETHENOD, however (see Abs. 2.1/2), utilises an expansion in powers of the current as a characteristic specification, a form which may prove useful in the analysis of certain cases.

BIBLIOGRAPHY

1. **Quantitative Relations in Detector Circuits :** B. LIEBOWITZ : *Proc. Inst. Radio Eng.*, 1917, **5**, 33-42.

This paper is primarily concerned with heterodyne reception, and gives a very comprehensive Fourier analysis of a discontinuous straight line and a curved (power series) characteristic.

The treatment is not entirely satisfactory from a practical point of view, since some essential factors (e.g., effect of output load) are left out of account.

(See also Abs. 2.3.2/3.)

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2. **On the Theory of Radio-telegraphic and Radio-telephonic Receiver Circuits :** J. BETHENOD : *Proc. Inst. Radio Eng.*, 1919, **7**, 517-25.

A theoretical analysis of heterodyne reception with a shunt detector. A simple type of characteristic (parabolic) is considered in relation to a particular type of circuit, the characteristic being represented by

$$v = ai + bi^2$$

The simple type of characteristic considered limits somewhat the applicability of the results arrived at. The inverse type of representation adopted is interesting, and has certain analytical advantages.

(See also Abs. 2.3.1/1.)

3. **The Rectifying Detector :** F. M. COLEBROOK : *Experimental Wireless*, 1925, **2**, 330-4, 394-403, 459-68.

This paper is mainly concerned with crystal detectors, but it gives a very general analysis of the rectifying characteristic, including the effect of the load, independently of any particular rectifier. A rectifier characteristic is specified in terms of a power series, and the rectifying action is explained analytically.

The conceptions of effective rectified voltage and effective rectifier internal resistance are introduced, and analytical expressions found for these quantities in terms of the constants of the assumed characteristic and the output circuit load.

The term " potential rectification " is also introduced, to signify rectification in which an infinite or very high impedance is presented to the continuous or modulation frequency output current.

A great part of the paper is concerned with the application of the above general ideas to typical crystal rectifier circuits. It is referred to again under that heading.

(See also Abs. 4.3/8.)

4. **A Theoretical and Experimental Investigation of Detection for Small Signals :** E. L. CHAFFEE and G. H. BROWNING : *Proc. Inst. Radio Eng.*, 1927, **15**, 113-53.

This paper is of importance as it gives a general analysis of a two-terminal rectifier in relation to pure and modulated continuous waves, taking account of the effect of the load. The authors make use of the conceptions of " rectification voltage " and rectifier internal resistance, introduced by COLEBROOK, and arrive at analytical expressions for these quantities in terms of the assumed characteristic of the rectifier.

Triode valve rectification is also considered on similar lines, and the paper is further referred to under that heading.

The practical application of the paper is limited to very small signals, since differential coefficients of higher order than the second are neglected.

(See also Abs. 2.3.2/24, 4.3/14.)

5. **Valve Detection :** P. DAVID : *C.R. Acad. Sci. Paris*, 1927, **184**, 1000-2.

The author proposes the use of " dynamic characteristics," i.e., curves showing the variation of rectified or continuous current with applied alternating (sine wave) electro-motive-force, as the proper basis for determining the behaviour of the rectifier under any given conditions. It is stated that this form of description is very suitable for practical cases of modulated continuous wave and heterodyne reception, in which cases only the first

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derivatives of the dynamic characteristic are required for the determination of the behaviour of the rectifier.

This paper is a very useful contribution to the general theory of rectification. See essay (2.1) for comment on term "dynamic characteristic."

(See also Abs. 2.3.2./21)

6. **Valve Detection :** P. DAVID : *Onde Élect.*, 1928, **7**, 313-50.

This paper amplifies the main thesis of the earlier paper (see above). Its detailed application to valve rectification is referred to under that heading.

(See also Abs. 2.3.2/21, 4.3/20.)

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2.2. RECTIFIERS OTHER THAN THERMIONIC VALVES

THE bibliography of this section is divided into two sub-sections 2.2.1 and 2.2.2. The first of these deals with the physical theory of crystal and other forms of contact detectors, and electrolytic detectors, and the second section with the application of such detectors to radio reception. The papers in Section 2.2.1 are not abstracted in any detail, since they are of very limited interest from the practical point of view of the present review of literature.

2.2.1. PHYSICAL THEORY

The work of PÉLABON (see Abs. 2.2.1/6, 7, 8, 10, 11) on this subject is, however, worthy of special mention, as it may have some bearing on the effects of unintentional imperfect contacts in receiver systems. PÉLABON has demonstrated experimentally, and has explained, on physical grounds, the fact that rectification takes place at an imperfect contact of two conductors, provided the thickness of the separating dielectric lies within certain limits (generally about 10^{-6} cm. apart). Such rectification can, for example, be obtained with a steel ball and a plane surface. In the case of galena, it is suggested that the pure galena itself is a moderately good conductor, and that the rectification is due to the insulation of the contact point by minute particles of free sulphur or other impurity.

Other papers included in the bibliography put forward alternative hypotheses as to the physical mechanism of contact detection, but no single hypothesis can be regarded as definitely established, though that of PÉLABON, referred to above, seems well supported by experimental and theoretical evidence.

Considerable practical interest attaches to the various forms of copper oxide film rectifying arrangements, but these are applied almost exclusively to the rectification of low audio-frequency alternating currents for accumulator charging or supply purposes, and are therefore not considered in any detail in this review. Such rectifiers seem at present unlikely to find effective application to signal detection.

Some interesting work has been carried out by H. ANDRÉ (see Abs. 2.2.1/21) on electrolytic rectification with an oxide coated metal electrode in a liquid colloidal metal suspension. He gives an elementary theory of the action of such systems, based on the partial penetration and subsequent reoxidising of the insulating

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oxide film. The application of such rectifiers is mainly to charging and supply purposes, but a brief account is given of a form applicable to signal detection. Its importance in this latter sphere is, however, very secondary.

2.2.2. Characteristics and Practical Application

As far as the application of contact detectors to radio reception is concerned, there is a surprising scarcity of recorded work on this subject. In fact, only one paper has been found which gives a comprehensive study of this subject, that by Colebrook (see Abs. 2.2.2/1) on "Rectifying Detectors." This paper gives a fairly complete survey of the static and operating characteristics of two of the most commonly used crystal detectors, namely, galena and perikon.

The behaviour of such detectors is interpreted as far as possible in terms of (1) input radio-frequency resistance; (2) output continuous or modulation-frequency rectified voltage; (3) effective internal resistance associated with the output voltage. Thus, writing E_c and R_c for the two latter quantities in the case of continuous wave rectification, the continuous current output with a load resistance R is given by

$$i_c = \frac{E_c}{R_c + R}$$

The principal conclusions arrived at are as follows:—

(1) The static and dynamic (see essay 2.1) characteristics of the detector appear to be similar but not identical; at radio-frequencies, however, the performance does not vary materially with frequency.

(2) The no load ($R = O$) value of R_c varies considerably with signal amplitude, tending to constancy at a hundred ohms or so for large signals (greater than half a volt) and rising to thousands of ohms for smaller signals.

(3) The no-load value of E_c is of the order of half the signal amplitude.

(4) The value of E_c increases with R, reaching an upper limit of 80 per cent. to 90 per cent. of the signal amplitude for fairly large signals (greater than half a volt) when R tends to infinity.

(5) The value of R_c does not vary greatly with R.

As far as the radio-frequency input resistance is concerned, it is shown that this decreases with signal amplitude and increases with R, being roughly proportional to R. Thus, for galena, the values found by harmonic analysis of a characteristic determined from the measured radio-frequency performance, showed a variation from 100 to 3,000 ohms for a variation of R from O to 3,000

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ohms. This is for the case in which the load is effectively short-circuited for radio-frequencies by a condenser. This variation of input resistance with load is shown to be due to the fact that the fall of direct potential in the load causes a change of the mean operating potential.

Curves showing the variation of transformation efficiency, i.e., ratio of output continuous current power to input radio-frequency power, are calculated on the above basis, and it is shown that the most favourable continuous current loads are fairly high (a few thousand ohms) in spite of the low effective internal output resistance of the detector.

The case of the rectification of modulated continuous waves is also considered, and the performance is interpreted in the same manner as for continuous wave rectification. It is shown that good quality reproduction of the modulation wave form should be obtainable with crystal detectors, the chief source of distortion being the variation of the electrical efficiency of the detector-telephone combination. From the latter point of view, and also from the point of view of overall transformation efficiency, moderately high resistance telephones would appear to be most suitable.

The output modulation frequency E.M.F. will in general be from 0·5 to 1 times the product of the modulation percentage and the carrier wave E.M.F. acting in the detector circuit. The corresponding internal resistance will be of the same order as the effective internal resistance for continuous wave rectification under similar load conditions.

Brief mention is also made of heterodyne reception. It is shown that the advantage obtainable as compared with " tikker " reception will vary roughly as the inverse of the signal E.M.F.

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2.2.1. Physical Theory

1. **Crystal Rectification of A.C.** : A. C. James : *Phil. Mag.*, 1925, **49**, 681-95.

2. **Galena Detectors** : J. Cayrel : *C.R. Acad. Sci. Paris*, 1925, **180**, 1728-31.

3. **Radio Detector Minerals** : E. Wherry : *J. Franklin Inst.*, 1925, **199**, 845.

4. **Detection Properties of Lead Dioxide** : J. Cayrel : *C.R. Acad. Sci. Paris*, 1925, **181**, 1127-8.

5. **Is Galena a Current Rectifier ?** : J. Cayrel : *Onde Élect.*, 1925, **4**, 155-7.

6. **Theory of Crystal Detection** : H. Pélabon : *C.R. Acad. Sci. Paris*, 1925, **181**, 776-8.

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7. **Detection by Metal Contacts :** H. PÉLABON : *Onde Élect.*, 1926, **5**, 464-75.

8. **On Detection :** H. PÉLABON : *Onde Élect.*, 1926, **5**, 141-9.

9. **The Rectifying Contact :** A. GUILLET : *C.R. Acad. Sci. Paris*, 1926, **183**, 350-2.

10. **Imperfect Contacts :** H. PÉLABON : *Onde Élect.*, 1927, **6**, 401-26.

11. **On Rectifying Contacts :** H. PÉLABON : *C.R. Acad. Sci. Paris*, 1927, **184**, 591-3.

12. **On a Verification of the Theory of Contact Detectors and Rectifiers :** M. DUBAR : *C.R. Acad. Sci. Paris*, 1927, **185**, 1023-4.

13. **The Direction of the Rectified Current in Crystal Detectors :** A. SCHLEEDE and H. BUGGISCH : *Phys. Z.*, 1927, **28**, 174-9.

14. **The Crystal Detector :** A. SCHLEEDE and H. BUGGISCH : *Jahrb. drahtl. Telegr.*, 1927, **30**, 190-3.

15. **Synthetic Galena Detectors and a new Theory of Crystal Rectifiers :** W. OGAWA, C. NEMOTO, S. KANEKO : *Elektrot. Laborat. Tôkyô Japan Researches*, 1927, No. 196, 77.

The above papers are concerned with theories and data of the physical mechanism of the crystal or contact rectification process, as distinct from the application of the phenomenon to wireless reception. They are not individually abstracted for that reason.

The papers by PÉLABON are mainly concerned with establishing, by analysis and measurement, the theory that contact rectification is due to the effect of the dielectric between two conducting surfaces in close proximity but not in contact, the two surfaces having different curvatures in the region of proximity.

The paper by SCHLEEDE and BUGGISCH gives a very useful critical résumé and bibliography of the physical theories of crystal detector action.

16. **The Johnson-Rahbek Contact effect as a Detector :** *Wireless Wld.*, 1926, **18**, 588.

17. **Rectification action of the so-called Electrostatic Relay, and an explanation of how Contact Detectors work :** W. KRAMER : *Z. f. Phys.*, 1924, **27**, 74-82.

These papers are concerned with rectification across surfaces of iron and agate in very intimate plane contact. The rectified currents obtained are too small for practical application. The action is attributed to ionic conduction in the agate and electronic conduction in the iron, resulting in a difference in the work required to liberate an electron from the two surfaces.

18. **A new type of Contact Rectifier :** L. GRONDAHL : *Phys. Rev.*, 1926, **27**, 813.

19. **High Pressure Powder Contact Rectifier :** LILIENFIELD and THOMAS : *Phys. Rev.*, 1927, **29**, 367.

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20. **The Dry Plate Rectifier :** H. FRÖHLICH : *Radio f. Alle.*, 1928, 165-70.

Descriptive articles dealing with the copper oxide film and similar dry rectifying surfaces, applicable to accumulator charging and supply voltages, but not directly to detection.

21. **Conductivity of Metallic Colloids and their Applications :** H. ANDRÉ : *Onde Élect.*, 1926, **5**, 28-46.

This paper describes rectifying arrangements using an oxide coated electrode immersed in a colloidal metal suspension. A theory of the action is given, but no experimental verification of the theory is recorded. Such rectifiers are intended mainly for battery charging and voltage supply purposes, but an experimental model for detection is also described. High efficiency is not claimed for the latter, and its characteristics are not given.

22. **Carborundum and its Rectification effect :** H. M. DOWSETT : *Radio Rev.*, 1921, **2**, 582-94, 642-59.

This paper describes a very detailed investigation of the carborundum detector and the effect of conditions of operation on the shape of the static characteristics. A theory of the action is put forward in the latter sections of the paper.

2.2.2. CHARACTERISTICS AND PRACTICAL APPLICATIONS

1. **The Rectifying Detector :** F. M. COLEBROOK : *Experimental Wireless*, 1925, **2**, 330-4, 394-403, 459-68.

(For Abstract, see essay on Section 2.2. and Abs. 2.1/3, 4.3/8.)

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2.3. VALVE RECTIFICATION

VALVE rectification can be divided into three main sections, covering respectively the application of diode, triode and tetrode valves to this purpose.

2.3.1. DIODE VALVE RECTIFICATION

As far as diodes are concerned there are only two items of any practical interest. The first is the "sodion" tube, developed and described by H. P. DONLE (see Abs. 2.3.1/2, 3, 4). This consists essentially of an evacuated glass bulb containing two electrodes, one of which has a U-shaped section and a heated filament, the latter being in proximity to the U-shaped electrode. The bulb contains a small quantity of sodium vapour, which is maintained at a suitable operating temperature by means of an external heating coil in series with the filament. The detector is used in much the same way as a crystal, and very high claims are made for it in respect of sensitivity, provided an input coupling appropriate to its relative low input impedance is used. It is stated to have pronounced frequency characteristics which can be modified to suit particular ranges of operation by modifications in construction.

In spite of the claims made for it, this detector does not seem to have come into general use.

The only other form of diode worthy of note is that constituted by a triode in which the grid is given a fixed positive potential with a view to neutralising the space charge. Such a rectifier will have an exceedingly high resistance in one direction and nearly straight line conductivity in the other. It is, of course, considerably less sensitive than a triode, particularly when the power absorption of the latter is annulled by means of retro-action, but it is stated that it will give a much purer reproduction of a modulation wave form. It has in fact been claimed that it will give distortionless reproduction, even with 100 per cent. modulation (see Abs. 2.3.1/9). It is probable that the arrangement will in fact, with suitable circuit conditions, give a nearly linear output with respect both to modulation frequency and depth, but a full analysis of the problem shows that it will not necessarily do so, even with a perfectly straight line static characteristic, unless the output load is a pure resistance practically independent of frequency. The full analysis does not appear to have been published, but a paper by the present author is in preparation.

A useful discussion of this type of rectifier is given by KIRKE (see Abs. 2.3.1/9), who appears to have originated the arrangement.

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He enumerates the precautions that have to be taken to utilise the desirable features of the diode. He also points out as a practical disadvantage the rather heavy drain on the grid battery, which may have to supply fifteen or twenty milliamperes continuously

A paper by BETHENOD, published in 1919 (see Abs. 2.3.1/1), gives a very complete analysis of heterodyne reception using a diode rectifier. The analysis is, in fact, almost too complete, inasmuch as it treats the aerial, aerial tuning coil, secondary, coupling, detector circuit, etc., as a single problem and attempts to generalise about it as a whole. This is a thoroughness which defeats its own ends, however, for it becomes difficult to isolate the essential variables of the problem and so make a systematic presentation of the matter. From the present point of view, however, the paper is of interest as suggesting, apparently for the first time, the conception of effective output rectifier voltage and internal resistance. In this respect the paper marks an advance in the systematic study of rectification.

Finally, as a matter of considerable theoretical interest in relation to valves in general, mention must be made of a paper by W. E. BENHAM (see Abs. 2.3.1/5), describing an investigation of the rectification efficiency of a diode (actually a triode with grid and anode connected) at very short wave-lengths (17 to 70 metres). The author claims to have demonstrated a frequency variation in performance at these wave-lengths, which implies a change in the dynamic characteristics of the diode. It would appear, however, that the method of measurement used for the purpose (a Dolazalek electrometer, the calibration of which was assumed to be independent of frequency), is open to criticism, and the conclusion cannot be considered as satisfactorily established in the absence of further confirmation.

2.3.2. TRIODE VALVE RECTIFICATION

It will be seen from the number of the abstracts that a very considerable volume of work has been published on triode valve rectification. The most important papers are those by E. B. MOULLIN and L. B. TURNER (see Abs. 2.3.2/8), F. M. COLEBROOK (see Abs. 2.3.2/15), P. DAVID (see Abs. 2.3.2/21) and S. BALLANTINE (see Abs. 2.3.2/29), and V. GROENEWALD, B. VAN DER POL, Junr., K. POSTHUMUS (see Abs. 2.3.2/20).

The most important single fact which emerges, and which finds general confirmation in the literature, is that the rectification performance of a triode valve can be predicted from the form of its static characteristics. That is to say, that as far as the valve itself is concerned there is no appreciable difference between its static and its dynamic characteristics. This can be regarded as established, at least up to a frequency of one or two million cycles per second. Beyond that point there is some conflict of evidence and

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further investigation is required. Unfortunately such an investigation must wait on the development of a satisfactory method of measuring alternating currents and potential differences at the very high frequencies involved.

Within the above limits of frequency, it would appear that in any future investigation of valve rectification the fullest possible use can be made of analytical methods, purely mathematical or graphical, based on easily determinable static characteristic data.

There is, however, a matter on which further experimental information is required, i.e., input impedance, as determined by inter-electrode capacity and circuit conditions. This factor, which will undoubtedly be of importance in many practical applications, has been almost entirely ignored in the hitherto published work. In fact only BALLANTINE considers the matter at all in detail, and then only in relation to a particular feature of one type of application. The neglect of this factor has led to a general misapprehension of the energy aspect of the processes in grid and anode circuit rectification. Time after time it is either explicitly stated or implicitly inferred that anode circuit rectification imposes no load on the input circuit, and that grid rectification necessarily will do so. In actual fact all that can be stated is that in grid rectification there is a certain part of the input impedance which is necessary for the rectification process. This may or may not be of importance compared with the remaining components of the input impedance. In certain cases it will probably be negligible compared with the other elements of the input impedance.

The published work on valve rectification is very varied in quality. There is a considerable degree of overlapping and repetition, probably attributable to a lack of systematic generalisation in the treatment of the subject. Further, there is a general failure to rate at its proper importance the energy transformation aspect of the rectification process. In most cases results are stated only in terms of input voltage and output current, a statement which gives a very limited view of the matter. COLEBROOK'S paper, it is true, refers to the energy conditions of rectification in some detail, but the treatment must be considered inadequate inasmuch as the input impedance due to inter-electrode capacity is ignored.

Considering the subject as a whole, the paper by P. DAVID (see Abs. 2.3.2/21) proposes a promising formulation of rectification processes in terms of what he calls " dynamic characteristics," i.e., a family of curves, derived experimentally or analytically, representing the continuous current output as a function of a mean steady voltage (e.g., grid voltage) and an applied sine wave voltage. The term " dynamic characteristic " is perhaps not well chosen, as it is more appropriate to the somewhat different sense described in section 2.1. The term " rectification characteristics " is a possible alternative. As shown in the abstract of this paper, the conception should prove of particular value in relation to modulated continuous

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wave rectification. In this connection the paper by BARCLAY, (see Abs. 2.3.2/22) should be useful in connection with the graphical determination of " rectification characteristics."

There is, in addition, the formulation proposed by COLEBROOK, namely, effective input impedance, effective rectified output voltage, and its associated effective internal resistance (or impedance). This would have the great advantage of specifying the energy conditions of the process.

Coming to more detail, it will be convenient to consider the subject under the two headings of (*a*) grid-circuit rectification, (*b*) anode-circuit rectification. Of the relation between the two it is enough to say that it is now understood that these processes give rise, in general, to anode current changes of opposite sense and should, therefore, not be allowed to occur simultaneously.

(*a*) *Grid-circuit rectification*

The early work on this subject was hampered by its association with the earlier forms of wireless transmission, i.e., by trains of damped oscillations. The term " grid-leak " is in fact a survival from those days and is quite inappropriate to the majority of present day applications of grid circuit rectification. COLEBROOK'S paper in 1925 appears to have been the first to emphasise explicitly the inappropriateness of the term and to show that the function of the " grid-leak " was that of a high resistance load in the rectifying circuit whereby the rectified voltage was made effective in controlling the anode current. The function of the grid condenser is simply to short circuit this load as far as the radio-frequency currents are concerned. In this connection it is stated by COLEBROOK that since the whole rectification process occurs in the grid circuit, there is no need for a by-pass condenser across the anode circuit modulation-frequency or continuous current load. This statement, however, is apt to be misleading. It is true so far as the actual rectification process is concerned, but it ignores the effect of the anode circuit load on the input impedance. In practice it will be found that in some cases such a by-pass condenser materially increases the output, due to the decrease in the effective input load produced thereby.

For the analysis of the process it is found that an exponential form

$$i_g = a\epsilon^{bv_g}$$

is appropriate for variations of v_g from about + 1 volt downwards to the negative limit. COLEBROOK gave a detailed analysis on these lines with a solution for continuous wave rectification in terms of what were subsequently pointed out to be Bessel functions. A very important factor, as far as modulated or heterodyned continuous waves are concerned, is the ratio of the modulation frequency to the carrier frequency. The process requires that the

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grid condenser shall, as far as possible, provide a path of negligible impedance at the carrier frequency and of very high impedance at the modulation frequency (relative, in both cases, to the grid-filament resistance). These conditions are mutually inconsistent except for a sufficiently high ratio of carrier to modulation frequency. Even at broadcast frequencies, a very considerable variation of output with modulation frequency may arise from this cause.

What is sometimes referred to as the " time constant " effect is actually no more than another way of presenting the above phenomenon. In connection with the reception of trains of damped waves, the time-constant mode of presentation is perhaps more appropriate, the phenomenon being of a transient rather than a quasi-stationary character.

There is, however, a further factor here which requires consideration where, as is usually the case, the operating voltage is a resonant potential difference in a tuned circuit. In such cases, the grid condenser controls the load imposed by the detector on the tuned circuit, and has an optimum value in consequence. In general this optimum value will be considerably smaller than the value appropriate to the case of a maintained input electro-motive-force, and the above amplitude distortion effect will be correspondingly reduced. The full practical significance of this has yet to be investigated, but a paper by COLEBROOK bearing on the subject appears in the *J. Inst. Elect. Eng.*, **67**, 157–169. (See Abs. 2.3.2/33.)

As far as the grid-circuit rectification of heterodyned continuous waves is concerned, this matter was investigated experimentally by ARMSTRONG in 1917, who showed that there was an optimum intensity for the local oscillation. However, it is clear from Armstrong's paper (and he himself points it out) that this is due not to the grid rectification process itself but to the fact that the lowering of the mean grid potential brings the lower curved region of the anode current characteristic into play, resulting in superimposed anode rectification. It was shown by E. V. APPLETON and MARY TAYLOR (see Abs. 2.3.2/14), that an optimum effect is possible apart from any anode curvature effect. Under practical conditions, however, such an optimum would probably be masked by the former effect.

The important frequency factor already mentioned in connection with the grid circuit rectification of modulated continuous waves plays an important part in connection with heterodyne reception. It was pointed out by E. V. APPLETON and MARY TAYLOR that the coupling condenser would have an optimum value on account of its beat frequency impedance, which acts as a shunt on the grid resistance and lowers the efficiency of the rectification process. COLEBROOK (see Abs. 2.3.2/15) emphasised the fact that this shunt effect would result in a very low efficiency in long wave heterodyne reception. In neither case, however, was the effect of the coupling

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condenser in the usual practical case of a resonant input circuit taken into account, and the matter must be considered to be incompletely investigated in this respect.

Finally, it has been pointed out by P. DAVID that certain apparent discrepancies between theory and practice in the grid rectification of medium and long heterodyned waves and in medium wave supersonic heterodyne, are to be accounted for by the fact that although the circuit is connected as for grid-circuit rectification it is in effect anode-circuit rectification which occurs. This matter is considered further under the heading of Anode Rectification.

(*b*) *Anode-circuit rectification*

The analysis of anode-circuit rectification is somewhat simpler than that of grid-circuit rectification, owing to the simpler form of the characteristic. In fact, for ordinary signal voltage amplitudes, the third and higher order derivatives of the characteristic can be neglected, leading to comparatively simple second degree equations for the operation. These are given in varying degrees of completeness in a large number of the papers mentioned in the abstract.

With anode-circuit rectification, it is necessary to provide a low-impedance capacity path for the radio-frequency currents, by means of a condenser connected across the anode circuit load. It has been pointed out by L. B. TURNER (see Abs. 2.3.2/19) that this fact introduces a possibility of amplitude-frequency distortion similar in nature to that associated with grid-circuit rectification. However, the magnitudes involved are such as to make the effect less pronounced, though it certainly occurs in certain cases of high resistance loads. This is a matter which requires more detailed investigation (see Abs. 2.3.2/33).

As applied to heterodyne reception, it was shown by E. V APPLETON and MARY TAYLOR (see Abs. 2.3.2/14) that if the anode-current characteristic can be represented by a power series up to the fifth power of the grid voltage, there will be an optimum local oscillator amplitude corresponding to the point of maximum slope of the continous rectified current characteristic. It appears, however, that this optimum will not in general be of a pronounced character. No curves have hitherto been published illustrating the nature and magnitude of the effect, but the general experience is that, judging by telephone intensity, nothing is to be gained by increasing the local oscillator amplitude beyond a certain point. That the optimum is not always realised in practice may of course be attributable to the fact that it is not sufficiently pronounced to be appreciated by ear, or that the local oscillator amplitudes available under usual conditions of reception cannot be made large enough to bring out the effect.

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(c) Comparison between grid and anode-circuit rectification

Grid-circuit rectification is inherently more sensitive than anode circuit rectification, at least as far as continuous wave rectification is concerned. For the reasons given above, the superiority will not be maintained in cases of modulated or heterodyned continuous wave rectification where the ratio of carrier to modulation frequency falls below a certain value, depending somewhat on the valve considered. In general, anode-circuit rectification should give a more faithful reproduction of telephony modulation, though here it must be remembered that the much higher internal anode circuit resistance associated with anode-circuit rectification imposes more severe requirements on the behaviour of the anode circuit load when the latter is part of an audio-frequency amplifying system. It is doubtful whether the overall superiority of anode-rectification is very marked, except perhaps in medium wave telephony (1,000 metres and over).

Comparison on an energy basis cannot be generalised owing to the complexity of the input impedance variations. In both cases, moreover, the input power absorption can be controlled over a wide range by suitable retro-action.

Some brief mention should be made of the "homodyne" or zero beat method of reception of telephony. This is only conveniently practicable in cases where the signal carrier wave is of sufficient intensity to "lock" the local oscillation at its own frequency. In such cases a modulation-frequency current is produced in the anode circuit of the local oscillator valve, with or without grid-leak and condenser, preferably without. It is a very sensitive method of reception but not one to be recommended as a practical policy, as the preliminary and subsequent adjustments of such a system are calculated to cause serious heterodyne interference, unless aerial radiation is prevented by using a coupling valve or by any other means.

A certain amount of work has been done on "soft" valves and valves containing traces of alkali vapour (see Abs. 2.3.2/13). The results described in the latter case present these valves in such a favourable light as rectifiers that a commercial development might reasonably have been anticipated. This, however, does not seem to have taken place in the four years since the work in question was published—at least as far as this country is concerned; though such valves may possibly be in use in America. Apart from the original paper, no information on the subject seems to be available.

In any further investigation of triode valve rectification the most useful objectives would seem to be :—

(*a*) A presentation, in as generalised a form as possible, of the rectification performance of a valve in terms of data easily obtainable from the static characteristics.

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(*b*) Correlation of performance with structural parameters for design purposes.

(*c*) Determination of appropriate circuit conditions of operation.

2.3.3. Tetrode Valve Rectification

Tetrode valves can be classified into three main groups: (*a*) double grid valves in which the inner grid is used mainly for the purpose of reducing the internal resistance of the valve; (*b*) valves containing an electrode which is used as a source of secondary emission; (*c*) valves in which the anode is internally screened from the grid (screen-grid valves).

With respect to ordinary double-grid valves, practically no detailed information has been published on the rectification performance, though a number of papers have appeared describing their operation in general terms in relation to certain particular applications, more especially heterodyne and supersonic-heterodyne reception. However, the normal rectification behaviour of such valves would probably not differ materially from that of triodes in any cases in which the interior grid was used as originally intended, namely, as a means of reducing the effect of the space charge and thus permitting the use of a much lower anode voltage.

With regard to certain special applications, more particularly frequency-changing arrangements such as supersonic heterodyne, it appears from the published literature that the possibilities have by no means been fully explored. In fact it may be said quite generally that the double-grid valve appears to deserve much more attention than has hitherto been given to it. Its development has, for some reason which is not at all clear, been neglected both by manufacturers and experimenters. It enjoyed a brief period of publicity in connection with the so-called " solodyne " circuit, and somewhat extravagant claims were made for it. It remains a fact, nevertheless, that the fundamental principle of the double-grid valve is both sound and of considerable practical importance, and it is to be hoped that the recent development of tetrode and pentode valves will induce designers and manufacturers to apply the experience so gained to the further development of this type. With the high emissivity filaments at present in use, there would seem to be no reason why valves with an internal resistance of hundreds rather than thousands of ohms should not be obtainable by means of the double grid principle. Even in the comparatively early days of valve design, the resistance had been brought as low as 6,000 ohms by this means.

No very hard and fast distinction can be drawn between the first and second groups of tetrode valves. In fact, the ordinary double-grid valves can be made to present negative resistance phenomena (see Abs. 2.3.3/7), though they are not particularly suitable for this purpose. For the most effective utilisation of

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secondary emission effects, a somewhat different design is required. Such a valve is the so-called " pliodynatron " described by HULL, HENNELLY and ELDER (see Abs 2.3.3/1). Here the outer grid is used as the anode, and the anode as a source of secondary emission, the inner grid exercising its usual control function. Such a valve appears to have very useful characteristics, more particularly in relation to heterodyne reception, where the local oscillation is produced through the agency of the negative resistance characteristic of the " dynode " or secondary emission electrode. For reasons which are not apparent in the above mentioned published account, this valve does not appear to have been developed commercially nor its practical possibilities fully explored. In this and some similar cases one would welcome the publication of reasons for the virtual disappearance of such arrangements from the field of applied science, if only as a guide in future investigation.

Comparatively little has so far been published on the use of screen-grid valves as rectifiers (see Abs. 2.3.3/20, 21). It appears that such valves are not very suitable for this purpose, the circuit conditions being somewhat critical and the high impedance loads required being unfavourable to modulation-frequency uniformity.

BIBLIOGRAPHY

2.3.1. DIODE VALVE RECTIFICATION

1. **On the Theory of Radio-telegraphic and Radio-telephonic Receiver Circuits :** J. BETHENOD : *Proc. Inst. Radio Eng.*, 1919, **7**, 517-25.

The paper gives an analysis of heterodyne reception with a diode rectifier, taking the effect of the telephone load into account, and considering the total receiving system as a whole. A parabolic characteristic is assumed for the rectifier.

The paper is to be commended as being one of the first to give due significance to the effect of the load. It suffers, however, from being over-complicated, due to insufficient separation of the problem into its essential parts. The conclusions arrived at are not directly applicable to modern practice.

(See also Abs. 2.1/2.)

2. **A new Non-interfering Detector :** H. P. DONLE : *Proc. Inst. Radio Eng.*, 1922, **11**, 97-109.

The paper describes a type of diode detector depending on conduction through ionised sodium vapour. An electrode consisting of liquid sodium is maintained at a suitable temperature by a heating coil external to the bulb of the detector. A second U-shaped electrode, referred to as the collector, is situated above the sodium and in fairly close proximity to an axial filament in series with the heater coil. The apparatus presents some peculiar features in performance, in particular a marked variation with radio frequency. This is claimed as an advantage in view of its possibilities in relation to selectivity. Very high voltage sensitivity is claimed. It is

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stated, however, that comparatively high input power is required for operation.

No information is given as to power efficiency, a rather important matter with this type of detector, since retro-action cannot be applied.

3. **Sodion Detector :** *Electrician*, 1923, **91**, 635.

This is a fairly full abstract of the paper by Donle referred to above.

4. **Applications of the " Sodion " Detector :** H. P. DONLE : *Proc. Inst. Radio Eng.*, 1924, **12**, 159-75.

The paper describes an improved form of the sodium vapour detector described in the former paper by the same author. In the new form of valve, a solid anode is used and the necessary sodium vapour is obtained by introducing a small quantity of sodium into the valve during the process of evacuation. A small heating coil is utilised, as in the earlier model. High sensitivity and purity of modulation-frequency reproduction are claimed. It is also claimed that the very loose aerial coupling appropriate to this detector gives the receiving system an unusually high degree of selectivity.

The grounds on which unusually high selectivity is claimed are not clear. It is certain that equally high selectivity could be obtained with any other detector by suitably adjusted circuit conditions. The " sodion " detector does not appear to have come into general use in this country.

5. **A Study of the Rectification Efficiencies of Thermionic Valves at Moderately High Frequencies :** W. E. BENHAM : *Phil. Mag.*, 1928, **5**, 323-34.

The paper records measurements on a diode (actually a triode with grid and anode connected) which point to a frequency variation in the rectification performance for frequencies ranging from 4 to 17 mega-cycles.

These observations appear to be at variance with those recorded by P. DAVID, who finds no material variation over a similar range of frequencies. It seems probable, however, that the accuracy of the measurements is not sufficient to justify any conclusions as to variation with frequency. BENHAM, for instance, used a Dolazalek electrometer as the input radio-frequency measuring instrument, and assumed that its calibration was independent of frequency.

(See also Abs. 4.3/19.)

6. **High Quality Reproduction :** R. P. G. DENMAN : *Wireless Wld.*, 1927, **20**, 92-100.

The above article describes the use of a triode valve as a diode rectifier, utilising the grid to neutralise the effect of the space charge and not as a control electrode. It is stated that by applying a sufficient positive potential to the grid and then using the anode-filament path for rectification, a practically perfect straight line rectifier can be obtained, and that with such a rectifier 100 per cent. modulation can be received without distortion.

Refer to essay, section 2.3, for comments on this type of rectifier.

7. **The Distortionless Diode :** H. F. SMITH : *Wireless Wld.*, 1928, **23**, 783-6.

This paper deals with the use of a triode valve with a positive grid potential of about 10 volts or so, as a diode rectifier. It is stated that such a rectifier will give " a rectified output strictly proportional to input voltage over a wide range." Appropriate circuit connections are shown, the detector being connected across a part only (about one-third) of the inductance of the input

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tuned circuit to reduce the damping caused by the detector load. The sensitivity of the arrangement is low compared with that of normal triode valve rectification, but it is claimed that the rectification is " distortionless," and that the use of the triode as a diode in this manner has the important advantage of separating the radio and audio-frequency stages of a receiver and thus minimising undesired inter-action effects. Practical details and circuit arrangements for broadcast reception are described.

The article is purely descriptive, and does not give any analytical or experimental confirmation of the statements respecting the uniformity and rectilinearity of the performance of this type of rectifier.

8. **Diode Rectifier :** P. G. DAVIDSON : *Wireless Wld.*, 1929, **24**, 23.

In a letter to the Editor, an interesting suggestion is made for obtaining straight line rectification by the use of two triodes of similar characteristics connected in series (filament of first to anode of second) with the grids biassed to give zero anode current. With an obvious notation—

$$i_a = f(v_{a1} + \mu v_{g1}) = f(v_{a2} + \mu v_{g2}) = 0.$$

For any *increase* δv_{g2} in the grid potential of the second valve—

$$i_a + \delta i_a = f(v_{a1} + \delta v_{a1} + \mu v_{g1}) = f(v_{a2} - \delta v_{a1} + \mu v_{g2} + \mu \delta v_{g2})$$

so that, whatever the nature of the function f

$$2\,\delta v_{a1} = \mu \delta v_{g2}$$

For a *decrease* in the grid potential of the second valve δv_{a1} will, of course, be zero. Thus it appears possible to obtain a perfectly linear relationship between a change in v_{a1} due to rectification and the applied alternating grid voltage. Further, the input resistance of such an arrangement should be considerably lower than that of a neutralised space charge diode.

The suggestion is an interesting one. The utilisation of the potential change without disturbing the linear conditions might present some difficulty. No actual output curves have been published.

9. **The Diode Rectifier :** H. L. KIRKE : *Wireless Wld.*, 1929, **24**, 32-35.

H. L. KIRKE, who originated the neutralised space-charge diode rectifier (the " Kirkifier "), gives here an account of the development of this type of detector. He states that an LS 5 B valve with a positive grid potential of 24 volts will give an output current which is linear up to about 2 m.a., and that this rectilinearity will ensure distortionless reproduction of modulation up to 1 m.a., even with 100 per cent. modulation. Further points are: The use of a pentode as the preceding valve (if a triode is used it should be neutrodyned) ; the grid bias battery has to supply rather large currents (up to 20 m.a. or so) ; a low L/C value for the tuned circuit is stated to be preferable to tapping down on the inductance, in order to provide a low impedance path for harmonics produced by the rectification process. (A letter from the Genl. Elect. Co. in the following number of the *Wireless World* calls attention to the heavy demand made on the filament of a triode valve used in this way as a diode.) The value of the separation of the radio from the audio-frequency stages given by the diode rectifier is emphasised, and it is suggested that a push-pull output system makes the most advantageous use of this separation.

It appears from the paper that the rectifying arrangement described will give an unusually faithful reproduction of modulation if the conditions of operation are carefully chosen. On the other hand, the numerous practical disadvantages of the arrangement will limit its general application. As a matter of theoretical interest, it should be pointed out that the rectilinearity of the continuous wave rectification

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characteristic is not in itself conclusive as evidence of distortionless reproduction of modulation, as the load conditions at audio-frequencies have to be taken into account.

10. **Performance of Valve Detectors :** W. B. MEDLAM and V. A. OSCHWALD : *J. Inst. Wireless Tech.*, 1929, **1**, 147-223.

The concluding section of this paper describes measurements made with a triode valve (voltage factor 10) used as a diode, the grid being maintained at a constant potential of 10 volts to neutralise the space charge. It is shown that with high resistance loads the bye-pass condenser may have an appreciable shunting effect at the higher audio-frequencies, leading to a decrease in the output voltage for a given input.

The results do not show the diode in a very favourable light, but it appears from a comparison with other publications on the subject that the type of valve used was not the most suitable for the purpose, and that the resistance load was too high to give the best performance from the point of view of rectilinearity. No curves are given of the variation of audio-frequency output with modulation percentage, nor of the continuous wave rectification characteristic with a resistance load, The latter would be considerably straighter than that with no load. which is given in the paper.

(See also Abs. 4.3/21.)

2.3.2. TRIODE VALVE RECTIFICATION

1. **The Working of the Three-electrode Vacuum Tube in Wireless Telegraphy :** G. VALLAURI : *Elettrotechnica*, 1917, **4**, 43-50, 63-70.

This paper describes what is probably one of the earliest attempts at a systematic analysis of the behaviour of a triode valve, or " audion " as it was then called, in its various applications. A limited study of a particular case of rectification (with a pure resistance anode circuit load) is given.

It appears from the description of the valves used that these differ so materially in characteristics from present day types as to render the work of historic interest only.

(See also Abs. 1.2.2/3.)

2. **A Study of Heterodyne Amplification by the Electron Relay :** E. H. ARMSTRONG : *Proc. Inst. Radio Eng.*, 1917, **5**, 145-59.

This describes an early experimental study of heterodyne detection. The results showed that there was an optimum local oscillator intensity for the actual conditions of the measurements, and that the gain in sensitivity under the optimum conditions, as compared with " equal heterodyne " reception, varied as some inverse power of the signal.

The paper was, in its time, a most valuable contribution to the subject of heterodyne reception. The conclusions would probably not be applicable in detail to present day practice, owing to the advance in valve design and technique generally.

(See also Abs. 4.3/1 and 1.2.4/1.)

3. **Quantitative Relations in Detector Circuits :** B. LIEBOWITZ : *Proc. Inst. Radio Eng.*, 1917, **5**, 33-42.

The paper gives a Fourier analysis of the behaviour of the so-called " perfect " rectifier (i.e., zero conductivity in one direction and constant

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conductivity in the other) and the " approximate " rectifier (i.e., zero conductivity in one direction and a curved variation of conductivity in the other).

The analysis is very detailed and thorough, and is useful for reference. The results arrived at, however, are of limited practical value, as the important effect of the output load is left out of account.

(See also Abs. 2.1/1.)

4. **Detection Coefficient of Thermionic Amplifiers :** H. J. VAN DER BIJL : *Phys. Rev.*, 1919, **13**, 311-14.

The paper describes an elementary investigation of a particular case of anode circuit rectification. It is assumed that the anode current change i due to a grid voltage change e can be expressed in the form

$$i = a_1 e + a_2 e^2,$$

and a_2 is determined by an audibility method (e being a modulated continuous wave) described in Section IV.

The effect of the load is not considered in detail and the conclusions are, therefore, of limited practical value, and refer only to the particular case considered and to small signal voltages.

(See also Abs. 4.3/2.)

5. **On the Detecting Efficiency of the Thermionic Detector :** H. J. VAN DER BIJL : *Proc. Inst. Radio Eng.*, 1919, **7**, 603-32.

This paper gives an analytical and experimental study of anode rectification of modulated continuous waves. A theoretical point of some interest is the proposition that for loads negligible compared with the internal resistance of the valve, the most suitable condition for anode bend rectification is defined by

$$v_a + \mu v_{vg} + \mu\epsilon = \mu v_f$$

where v_f is the filament voltage and ϵ a small constant.

This result does not appear elsewhere. It would apparently not apply for appreciable anode circuit loads. A somewhat over simplified representation of the rectification process is adopted, and the significance of an appropriate anode circuit load is not considered. These facts detract from the value of the results obtained and from the suggested method of comparing valves in respect of rectification. The method of measurement is described in Section IV.

6. **A Theoretical Study of the Three-element Vacuum Tube :** J. R. CARSON : *Proc. Inst. Radio Eng.*, 1919, **7**, 187-200.

This paper is characterised by the analytical thoroughness which is always found in Carson's work. It investigates the anode circuit rectification of a continuous wave of any wave form representable by a Fourier series, with any given impedance in the anode circuit, the anode current characteristic being represented by an unrestricted power series. Very complete but complicated formulæ are arrived at for the various components of the anode current.

The paper would have been more valuable from a practical point of view if the application to design had been more completely discussed. As it is, the practical implications are left unexpressed and cannot, therefore, be abstracted.

(See also Abs. 1.2.2/8.)

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7. **Efficiency of a Single Thermionic Valve as a Detector :** E. O. HULBURT and G. BREIT : *Jahrb. drahtl. Telegr.*, 1921, **17**, 119-32.

The paper assumes a general representation

$$i_a = f(v_a, v_g)$$

for the anode current of a valve, and calculates in terms of the first two partial differential coefficients the value of the so-called anode circuit detection coefficient

$$lim._{E=0} i_0/E^2$$

where i_0 is the change of anode current corresponding to a signal of R.M.S. value E acting on the grid. The variation of the coefficient with the impedance of a tuned circuit in the anode was investigated and showed qualitative agreement with the theory. The differences between calculated and measured values are attributed to inter-electrode capacity effects.

The conclusions are of limited practical importance, the circuit studied being one not in general use. The method of measurement is described in Section IV.

(See also Abs. 4.3/3.)

8. **The Thermionic Triode as Rectifier :** E. B. MOULLIN and L. B. TURNER : *J. Inst. Elect. Eng.*, 1922, **60**, 706-19.

This paper is an important contribution to the systematic study of valve rectification, and covers the anode and grid circuit rectification of pure continuous waves, heterodyne modulated continuous waves and damped wave trains.

In the first place, the analysis of the no-load anode circuit rectification of a pure continuous wave was confirmed experimentally by measurements with a low audio frequency supply (90 cycles per second), and it was shown that for signals up to about 1·5 volts (R.M.S.) a parabolic characteristic could be assumed. In a similar manner it was shown that the grid-current characteristic is more nearly exponential than parabolic.

Grid circuit rectification was then analysed on the basis of a parabolic form of characteristic for small signals and the analysis confirmed experimentally. The suitable values of grid condenser and "grid leak" were discussed.

The treatment of grid and anode circuit heterodyne rectification is purely theoretical. The advantage of employing a large local oscillator amplitude in the former case is illustrated by calculation, and it is shown that it is the slope rather than the ordinate of the rectified current curve that is important in this connection.

The paper concludes with an approximate discussion of the effect of the load impedance.

The paper was an exceedingly useful one as far as contemporary conditions were concerned, and gave data of considerable value in relation to the valves actually used. This data would probably require revision in relation to modern valves. The treatment of grid-circuit heterodyne rectification cannot, however, be considered entirely satisfactory, as insufficient attention is paid to the effect of the audio-frequency impedance of the condenser-leak combination.

(See also Abs. 4.3/4.)

9. **Detecting Efficiency of the Resistance-Capacity Coupled Amplifier to 6,000 metres:** W. G. BROMBACHER : *Phys. Rev.*, 1922, **20**, 433-40.

The paper deals with the continuous wave anode-circuit rectification performance of a two-valve amplifier-detector combination, and the objective

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of the work was the confirmation of a very complex formula applicable to this particular type of combination.

The agreement obtained is not very close (10 per cent. or so). Practically no attempt is made to analyse the data in relation to the various causes operating, and very little information of general applicability is obtained. It appears to be a careful and thorough set of measurements, yielding a poor return on account of the rather ineffective design and analysis of the experiments, since it treats as a single problem what is, in fact, a group of problems.

See also Abs.4.3/6.)

10. **Some Remarks on the Working of a Detector Valve :** E. FROMY : *Onde Élect.*, 1923, **2**, 411-19.

The author gives an elementary qualitative discussion of grid and anode-circuit rectification, pointing out that these may oppose or reinforce each other according as grid-circuit rectification is combined with lower or upper bend anode-circuit rectification. He also shows that the time-lag associated with grid-circuit rectification on account of the grid condenser may give rise to anomalous and undesirable results with ill chosen values of grid condenser.

11. **Study of a Process of Reception by a Modulated Valve :** E. FROMY : *Onde Élect.*, 1923, **2**, 168-73.

The author describes a novel valve detector circuit applicable to telephony and telegraphy reception at wave-lengths less than 1,000 metres. It consists of a tuned grid circuit oscillator, the tuned circuit being connected to the grid through a large variable capacity of 0·01 μF., shunted by a resistance of half a megohm or less. The so-called " squegging " condition is set up (i.e., regularly interrupted radio-frequency oscillations), the interruption being at a high audio-frequency, and the incoming signal detected by a change in the pitch of the interruption note. For telephony reception, the interruption frequency is made very high and " by-passed " for the telephone by a filter, and reception is then made by means of anode current changes due to the signal.

The advantages claimed are slightly higher than normal sensitivity, with complete stability of operation.

12. **An Experimental Study of Processes of Detection of High Frequency Oscillations :** R. DUBOIS : *Onde Élect.*, 1924, **3**, 285-98, 347-63.

This paper, after a qualitative discussion of the phenomena of rectification by crystal and grid and anode-circuit curvature, gives an oscillographic study of the modulation wave-form produced by the rectification of a " chopped " continuous wave. It is shown that anode-circuit and crystal rectification will give a faithful reproduction of the wave form, but that grid-circuit rectification will not, owing to the time-constant limitations of the grid-condenser and grid leak combination.

The paper is thus a confirmation of a fact now well appreciated in this connection, but the very special type of modulation wave form considered and the special conditions of measurement make it difficult to draw any generalised practical conclusions from the work.

(See also Abs. 4.3/7.)

13. **Alkali Vapour Detector Tubes :** H. A. BROWN and C. T. KNIPP: *J. Amer. Inst. Elect. Eng.*, 1924, **43**, 26-32.

This paper describes an investigation of the characteristics, more particularly as detectors, of triode valves containing vapour of a sodium-

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potassium alloy. Details of construction and design are given, together with numerous characteristics and anode-rectification curves. It is claimed that such values are more sensitive as detectors than hard triodes and less " critical " and more " stable " in operation.

The meaning of these terms is not made quite clear. Such valves do not appear to have come into general use.

14. **Optimum Heterodyne Reception :** E. V. APPLETON and MARY TAYLOR : *Proc. Inst. Radio Eng.*, 1924, **12**, 277-93.

The authors show by a theoretical analysis that if a power series continued up to the fifth power of the grid voltage represents the anode current variation in a triode, there will be an optimum local oscillator amplitude in the case of anode circuit heterodyne rectification, this optimum value being associated with the point of maximum slope of the curve of continuous rectified current.

The paper further gives a very interesting analysis of grid-circuit rectification. It is shown that in this case an optimum local oscillator amplitude may arise from the considerations already mentioned, but that another optimum condition will arise in this case from the opposition of grid and anode-circuit rectification.

The function of the grid coupling condenser is considered in some detail, and it is shown that, for a given input voltage, the condenser will have an optimum value, this arising from the fact that it is required to have a high impedance at the beat frequency and a negligible impedance at the radio-frequency (relative to the effective resistance of the grid-filament path).

The following general conclusions are arrived at in the paper :—

(*a*) The optimum value of the local oscillation amplitude is independent of the received signal magnitude when the latter is small.

(*b*) The magnitude of the combination tone is directly proportional to the strength of the signal oscillation when the local oscillation has been adjusted for optimum heterodyne.

(*c*) The ratio of the optimum value of the combination tone to that produced with equal heterodyne is inversely proportional to the strength of the received signal.

The value of the work would have been enhanced by an experimental curve showing the actual variation of the beat-frequency output, as the paper gives no indication of the sharpness of this optimum. It would also have been desirable to indicate to what extent the conclusion would be valid when the effect of the output load is taken into account.

In connection with the discussion of the magnitude of the coupling condenser in grid rectification, it should be noted that in most practical cases of reception the input voltage is a resonant potential difference across a tuned circuit, and that the analysis given will not necessarily apply to such cases, since the load imposed by the rectifying circuit will also depend on the magnitude of the coupling condenser. The recent work of P. DAVID should also be consulted in relation to this matter.

15. **The Rectification of Small Radio-frequency Potential Differences by means of Triode Valves :** F. M. COLEBROOK : *Experimental Wireless*, 1925, **2**, 865-77, 946-57 ; 1926, **3**, 34-40, 90-93.

This paper gives a systematic theoretical and experimental investigation of grid and anode-circuit rectification of pure, telephony-modulated and heterodyne-modulated continuous waves.

Grid circuit rectification is analysed on the basis of an exponential representation of the grid-current characteristic, which is confirmed by the measurement of typical characteristics. It is shown that the function of the grid-leak is to make externally operative, so to speak, the internal rectified E.M.F. produced by the action of the curved characteristic on an applied

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alternating E.M.F., the grid condenser having the auxiliary rôle of short circuiting the grid-leak for the radio-frequency components of the current.

The process of continuous wave rectification is analysed and the analysis is confirmed experimentally. Appropriate values of grid-leak and grid condenser are determined analytically and confirmed by measurement. The effective input resistance is also calculated in terms of the characteristic.

On the basis of experimentally verified assumptions, it is shown that the grid-circuit rectification of telephony modulated continuous waves may, and in many cases will, give rise to an appreciable degree of amplitude-frequency distortion, due to the variation of the audio-frequency impedance of the grid-condenser, grid-leak combination. This is the quasi-stationary aspect of the time constant effect in the reception of spark generated waves. As a corollary of the above it is shown that grid-circuit rectification is likely to be ineffective in long and medium-wave heterodyne reception.

Anode-circuit rectification is analysed on the basis of the assumption of a parabolic form for the characteristic. It is shown that this process is inherently less sensitive than grid-rectification.

It is pointed out that anode-rectification is free from the inherent frequency distortion involved in grid-circuit rectification. Anode-circuit heterodyne rectification is considered, and it is shown that in this connection the important factor is not the slope of the rectified continuous current-signal voltage line, as in telephony reception, but that of the rectified continuous current—(signal voltage)^2 line. This statement, however, though mathematically justifiable, is apt to be misleading. If the signal voltage is small compared with the local oscillator voltage, it is the slope of the rectified current —signal voltage line which is important, as pointed qut by MOULLIN and TURNER.

Brief mention is made of homodyne reception, and the paper concludes with a comparison in general terms of crystal and valve rectification.

It must be noted that the paper refers throughout to operation with a maintained input E.M.F. Thus the analysis of the coupling condenser value would require modification in the practical case of a resonant input potential difference (see essay 2.3.). A further limitation in the paper is that no account is taken of inter-electrode capacities, and the coupling effects of these. Thus the input resistance calculations refer only to that component of the input resistance which is due to the rectification process. This omission will also modify the coupling capacity analysis.

(See also Abs. 4.3/12.)

16. **Detecting Characteristics of Electron Tubes :** H. M. FREEMAN : *Proc. Inst. Radio Eng.*, 1925, **13**, 611-23.

The object of the author is the study of the rectification performance of a valve in relation to its constructional constants. The paper describes certain measurements which show a qualitative agreement between Morecroft's formula for grid-circuit rectification and the measured results with certain particular valves.

No systematic analysis of the actual experimental conditions are given, and those few conclusions which can be drawn from the experimental curves are applicable only to the particular valves and the particular measuring circuit employed.

It is interesting to note that the measurements were made in a room completely enclosed in wire netting, for screening against outside interference.

(See also Abs. 4.3/11.)

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17. **Operation of Thermionic Vacuum Tube Circuits :** F. B. LLEWELLYN : *Bell System Techn. J.*, 1926, **5**, 433-62.

The author gives a very thorough analysis of the behaviour of a triode valve with generalised impedances connected between grid and filament, anode and filament, based upon the generalised representation

$$i_a = i_a\,(E_g,\ E_p)$$
$$i_g = i_g\,(E_g,\ E_p)$$

This does not assume the constancy of μ, nor that E_p is independent of i_a. A new valve characteristic $v\ (= dE_p/dE_g$ for a constant $i_g)$ is introduced, and is called the reflex factor. It is analogous in the grid circuit to μ in the anode circuit.

It is shown that in certain cases the variation of μ may play an appreciable part, particularly in rectification with high anode-circuit loads.

Anode and grid-circuit rectification are discussed in terms of the completely generalised representation of the valve, and taking account of the inter-electrode capacities. It is shown that in general, types of rectification will occur simultaneously to some degree. The value of the grid condenser is considered, and it is shown that it will have an optimum value for any given modulation and carrier frequency.

The formulæ arrived at are of necessity very complicated in form on account of the high degree of generalisation arrived at. The usual formulæ are arrived at as special cases on making the usual assumptions as to the constancy of μ, etc. It does not appear that the additional complications involved in the full generalisation are really justified by any practical consideration, but the formulæ obtained are of value from a purely theoretical point of view.

(See also Abs. 1.2.2/17.)

18. **Theory of Detection in a High Vacuum Thermionic Tube :** L. SMITH : *Proc. Inst. Radio Eng.*, 1926, **14**, 649-62.

This is a theoretical discussion of grid-circuit rectification on conventional lines and considers in particular the effect, already pointed out by various writers, of the grid condenser on the audio-frequency output. It is shown that in the graphical representation of grid-circuit rectification of a modulated continuous wave, the representative point follows a closed path of more or less elliptical shape.

The author's summary states that new ideas are put forward, but the paper does not in fact contain anything new, apart from the demonstration of the elliptical locus of the representative point, a matter of theoretical rather than of practical interest.

19. **Notes on Wireless Matters :** L. B. TURNER : *Electrician*, 1926, **97**, 288-89, 412, 420.

This is a somewhat elementary theoretical account of grid and anode-circuit rectification. With regard to the latter, it is stated that " anode-circuit " rectification need take no power from the radio-frequency source.

In considering grid-circuit rectification, the writer tries to show by a simple theoretical discussion that it is not materially worse than any of the alternative methods in the matter of frequency distortion.

It is shown that with both types of rectifier (grid and anode-circuit), and also with a crystal, a large output circuit resistance tends to give a linear output characteristic, particularly for fairly large input voltages.

The statement with regard to the power absorption with anode rectification is somewhat misleading, as the input impedance due to inter-electrode capacity coupling is left out of account. In the comparison of grid and anode-circuit rectification from the point of view of

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modulation frequency uniformity, it would probably be more true to say that conditions for uniformity are more easily fulfilled in anode than in grid-circuit rectification.

20. **Grid Rectification:** V. B. F. GROENEVELD, B. VAN DER POL, JR., and K. POSTHUMUS: *Jahrb. drahtl. Telegr.*, 1927, **29**, 139-47.

This is a useful contribution to the theory of grid rectification. An exponential grid current characteristic is assumed and is shown to be a consequence of the physical theory of operation of the valve. The term "temperature voltage" (Temperatur spannung) is introduced to indicate that grid voltage increase which gives an e-fold increase of grid current, and it is shown that the rectification performance for small signals, as measured by the change of mean grid voltage, is the same for all valves for signal voltages having the same ratio to the "temperatur spannung." For large signals the grid voltage change tends to equality with the signal amplitude for all valves.

(See also Abs. 4.3/16.)

21. **Valve Detection:** P. DAVID: *C.R. Acad. Sci. Paris*, 1927, **184**, 1000-2 and *Onde Élect.*, 1928, **7**, 313-59.

These papers give a very useful generalisation of the problem of rectification, with particular reference to triode valves. It proposes the specification of a rectifying (non-linear) characteristic,

$$i = f(v)$$

by means of a family of so-called dynamic characteristics

$$i_c = \phi(v_0, \hat{e})$$

which record the continuous current corresponding to a mean voltage v_0 and an alternating amplitude $\hat{e}$. While these will not, in general, admit of any simple algebraic formulation, it is suggested that the numerous practical cases in which the output corresponding to a small change $\delta\hat{e}$ of a relatively large local oscillation or carrier wave $\hat{e}$ can be given a simple algebraic formulation in terms of the partial derivatives.

$$\delta\phi/\delta v_0 \text{ and } \delta\phi/\delta\hat{e}.$$

The *Comptes Rendus* paper gives a useful comparative table for different types of rectification based on these ideas.

The paper in *L'Onde Electrique* gives examples of the practical application of these ideas. It shows that over certain limited ranges f can be given simple algebraic forms and the ϕ curves determined by analysis, but that no single form is in general completely comprehensive, though the parabolic form for anode rectification and the exponential form for grid rectification have a useful range of validity.

In the discussion of grid rectification, the dependence of the efficiency of the process on a high ratio of radio to modulation frequency is emphasised, and it is shown to be very unsuited to long wave heterodyne and medium wave superheterodyne reception. It is shown, in fact, that in such cases anode-circuit rectification is probably arrived at by suitable adjustment of the local oscillator amplitude, in spite of the connection of the circuit as for grid rectification.

Another feature of the paper of considerable theoretical interest is the extension of the measurement of rectification performance down to 14 metres, and it is shown that within somewhat wide limits of experimental error a valve does not show any pronounced frequency effects even under these extreme conditions.

(See also Abs. 2.1/5, 4.3/20.)

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22. **Numerical Solution of Grid Rectification Problems :** W. A. BARCLAY : *Experimental Wireless*, 1927, **4**, 459-66, 552-8.

This paper describes a simple and apparently very effective method (to which the author gives the name " three co-ordinate method ") of determining the " dynamic " or rectification characteristics from the static characteristic in the case of grid rectification. The method depends on the proposition

$$\int_0^{\pi} f(\cos\theta)\, d\theta = \frac{\pi}{p}\left\{ f\left(\cos\frac{\pi}{2p}\right) + f\left(\cos\frac{3\pi}{2p}\right) + \ldots\ldots f\left(\cos\frac{\overline{2p-1}\pi}{p}\right)\right\}$$

which is true if the $2pth$ differences of f are zero. For application to a grid current characteristic it can be assumed that $p = 3$.

The paper deals entirely with the application of this proposition to grid current characteristics, and does not give any detailed analysis derived therefrom of the various types of rectification process. It is a useful complement of the paper by DAVID.

23. **Note on Detection by Grid Condenser and Leak :** W. VAN B. ROBERTS : *Proc. Inst. Radio Eng.*, 1927, **15**, 793-6.

The paper gives the result of an analysis of the frequency distortion in grid rectification in a form which is identical with that obtained by COLEBROOK. It is stated, however, that the frequency distortion effect is negligible.

This conclusion appears to be based on a particular set of values, including a figure 60,000 ohms for the alternating current resistance of the grid-filament path. This is an unusually low value to assign to this resistance.

24. **A Theoretical and Experimental Investigation of Detection for Small Signals :** E. L. CHAFFEE and G. H. BROWNING : *Proc. Inst. Radio Eng.*, 1927, **15**, 113-53.

The paper is comprehensive in scope and considers grid and anode-circuit rectification of continuous and modulated continuous waves, and in addition deals briefly with the effect of retro-action. The treatment is limited to signals so small that the effect of third and higher derivatives of the characteristics can be neglected. The conceptions of effective output rectified voltage and associated internal resistance are made use of, and expressions are found for these in terms of the assumed simple forms of characteristics.

Grid-circuit rectification is consideredi n detail and the variation with modulation-frequency is pointed out and illustrated by calculated curves.

In connection with the rectification of modulated continuous waves, it is shown that any rectification process will give rise to the production of sum and difference tones when more than one modulation frequency is present, and that those can only be kept small relative to the original modulation-frequencies by keeping the modulation percentage small. This is a useful general conclusion.

Experimental verification of the analytical treatment is obtained by the null method described in Section IV. Much of the measurement refers, however, to soft valves, and the corresponding curves show the usual irregularities of form associated with such valves.

The treatment throughout is very thorough and systematic within the limitations already mentioned, and expressed in a somewhat elaborate system of notation specially invented by one of the authors for the purpose of the analysis of circuit and valve problems.

The question of the optimum adjustment of the grid capacity in the case of the rectification of a resonant potential difference is not considered, and the treatment of frequency distortion must be considered incomplete to this extent.

(See also Abs. 4.3/14, 2.1/4.)

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25. **Anode Rectification :** M. v. ARDENNE : *Jahrb. drahtl. Telegr.*, 1927, **29**, 82-88.

The paper describes a theoretical and experimental investigation of anode-circuit rectification with various anode circuit load conditions. The analysis assumes a parabolic form for the characteristic over the range of the potential changes involved, and arrives at the approximate formulæ

$$\delta i_a = \left(\frac{R_a}{R + R_a}\right)^3 \frac{\delta^2 i_a}{\delta v_g^2} \frac{E_g^2}{4}$$

and

$$\delta i_a = \left(\frac{R_a}{R + R_a}\right) \frac{\delta^2 i_a}{\delta v_g^2} \frac{E_g^2}{4}$$

where, in the first case, the anode circuit load is the pure resistance R, and in the second, this resistance is shunted by a condenser of negligible reactance at the frequency considered. (At high radio-frequencies of the order of 10^6, the anode-filament capacity of the valve may be sufficient for this purpose.)

The deductions are confirmed qualitatively by measurements made at an audio-frequency (800 cycles).

It is shown that an inductive load of high impedance and low resistance at the frequency of operation will give a greatly reduced rectification effect owing to the resulting straightness of the dynamic characteristic as compared with the static characteristic.

(See also Abs. 4.3/13.)

26. **Voltage Detection Coefficient :** E. L. CHAFFEE : *Proc. Inst. Radio Eng.*, 1927, **15**, 946-57.

This paper is a continuation of the work previously described by the same author, and gives the results of measurements of the effective output audio-frequency voltage given by the grid and anode-circuit rectification of a pure tone modulated continuous wave. The method of measurement is described in Section IV.

An elementary account is also given of the effect of the input impedance of a rectifier valve on a tuned circuit. It appears to be assumed that only in the case of grid rectification is this a factor of any importance.

The results obtained are only applicable to the particular cases considered. The only general conclusion arrived at is that a high voltage factor is generally advantageous for rectification.

The apparent assumption that the input load is only of importance in the case of grid rectification is incorrect as a generalisation.

(See also Abs. 4.3/15.)

27. **A Novel Valve Detector :** H. J. NEILL : *Experimental Wireless*, 1928, **5**, 74-76.

The scheme described by the author consists of two valves connected as for push-pull operation, with the load in the common high-tension lead. It is shown that if two such valves are used as rectifiers, the filament being connected to the centre of the input coil and the two grids to the ends, currents of carrier frequency will not be present in the output circuit, and are thus prevented from getting into the audio-frequency stages.

The scheme certainly has valuable possibilities, but it should be pointed out that the multiple carrier frequency currents due to the rectification process are not suppressed by this connection. These, however, are usually easy to eliminate as a source of trouble on account of their high radio frequency.

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28. **Anode Rectification :** M. v. ARDENNE : *Jahrb. drahtl. Telegr.*, 1928, **31**, 51.

This is an addendum to the above mentioned paper by the same author. On the basis of the formulæ already quoted, the ratio of anode voltage change ($R\delta i_a$) to input R.M.S. radio-frequency grid voltage (E_g) is expressed in the form

$$\frac{R\delta i_a}{E_g} = \frac{\mu R}{R+R_a} \cdot \frac{E_g}{4} \cdot \frac{\delta^2 i_a}{\delta v_g^2} \frac{\delta i_a}{\delta v_g}$$

for the case in which R is assumed to be short-circuited to the radio-frequency currents by a suitable capacity.

It is stated on the basis of this formula that the maximum sensitivity is obtainable by using valves of very high voltage factor with a very high value of external resistance, since the first and third factors are both large under these conditions.

The author does not give experimental results in support of this conclusion, and the low audio-frequency response characteristic of the arrangement proposed is not considered.

29. **Detection by Grid Rectification with the High Vacuum Triode :** S. BALLANTINE : *Proc. Inst. Radio Eng.*, 1928, **16**, 593-613.

In the analysis of the grid rectification of a modulated continuous wave, an expression is found for the modulation-frequency change of grid potential. This contains a factor similar in form to that introduced by COLEBROOK to represent the frequency distortion given by this kind of rectification, but more general in applicability as it takes account of the complete input admittance of the detector valve, including the effect of the anode-circuit load. The factor is written in the generalised vector form

$$\frac{Z}{1+ZY_0}$$

where Z is the modulation-frequency impedance of the grid condenser and grid-leak, and Y_0 the input modulation-frequency admittance of the valve.

It could have been put in the even simpler form

$$\frac{1}{Y+Y_0}$$

by using the admittance Y instead of its reciprocal Z.

It is shown, however, by means of measured values of Y_0 corresponding to typical anode-circuit loads, that the main effect is due to the actual grid-filament conductivity. The magnitude of the frequency effect is illustrated in typical cases. Incidentally, the writer makes the same criticism of Roberts' paper on this aspect of the process that is made in the present résumé, namely, that Roberts' assumed values of effective grid-filament conductivity were quite misleading.

An ingenious method of measuring the audio-frequency distortion effect by applying an audio-frequency voltage direct in the rectifying circuit is described. Measured results show that the grid-rectification frequency distortion is comparable in importance with the other sources of distortion in a receiver. It is also shown that anode-circuit rectification is free from this particular defect.

It should be pointed out, however, that Ballantine's treatment of this subject is open to the criticism that applies to all hitherto published analyses, namely, that the significance of the grid condenser as a means of controlling the load of the rectifier on the tuned circuit is overlooked. This will in most cases lead to a much smaller optimum condenser value than is usually employed, resulting in a correspondingly smaller frequency distortion effect.

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The treatment as a whole is limited to signals so small that a parabolic form can be assumed for the characteristic. Within these limits, however, the paper is a very good example of the systematic and generalised treatment of its subject.

It is interesting to note that BALLANTINE arrives at the same conclusion as DAVID with respect to supersonic-heterodyne rectification, namely, that in all probability it is invariably anode-circuit rectification that takes place.

(See also Abs. 4.3/17.)

30. **Direct Coupled Detector and Amplifiers with Automatic Grid Bias :** E. H. LOFTIN and S. YOUNG WHITE : *Proc. Inst. Radio Eng.*, 1928, **16,** 281-6.

A novel detector amplifying system which appears to have interesting possibilities is described. A typical combination is as follows. The grid of the second valve is connected through a high resistance (value not stated, but apparently of the order of 0·25 megohm) to the positive end of the filament, and also through a battery of four volts or so to the anode of the first valve. The signal is applied to the grid of the first valve. The voltage on the grid of the second valve depends on the anode-filament resistance of the first valve, which is, of course, varied by the signal. A similar coupling can be employed between the second and a further amplifying valve, and so on. It is apparently necessary to employ some means whereby the anode current in the last valve controls the grid bias of the first, to prevent a saturation condition being set up by a large carrier wave intensity.

The description is in rather vague terms, with an absence of detailed explanation or specification. The circuit is, however, an interesting suggestion, and it is claimed that it is free from microphonic or retro-active effects. The coupling would seem to have a uniform frequency characteristic, but self capacities and input capacities may have undesired effects. This is not considered in the paper.

31. **Fading and the Detector :** *Wireless Wld.*, 1929, **23,** 485.

It is suggested in this short unsigned article that grid-circuit rectification will tend to minimise the variation of signal strength caused by fading, since it is more sensitive for weak signals than for strong signals. With anode circuit rectification, on the other hand, the reverse is the case.

32. **Some Principles of Grid-leak Grid-condenser Detection :** F. E. TERMAN : *Proc. Inst. Radio Eng.*, 1929, **16,** 1384-97.

The paper deals with grid-circuit rectification by the usual method, with particular reference to modulated input voltage. The process is interpreted in terms of an equivalent rectified voltage acting in the grid circuit. The author arrived independently at this form of representation, but states that it had been previously proposed by CHAFFEE, COLEBROOK and CARSON, CARSON being apparently the first to introduce the idea, though not in its most general form.

Comparatively simple formulæ for the effective rectified voltages are given, e.g., with a modulated E.M.F. $\hat{e}\ (1 + \hat{m} \sin nt)$ the amplitude of the effective modulation frequency voltage acting in the grid circuit is given as $\hat{e}^2\ \hat{m}/v$, where $v = 2\ R_g/(\delta R_c/\delta e_g)$ and $R_g = \delta e_g/\delta i_g$, e_g being the mean grid potential under operating conditions. This and the other simple formulæ given apply only to very small signal voltages (·05 volt or less). A simple bridge arrangement for measuring v and R_g is described. The effect of the magnitude of the grid condenser on sensitivity and modulation frequency distortion is considered on similar lines to those followed by COLEBROOK and BALLANTINE (see Abs. 2.3.2/15 and 2.3.2/29).

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The paper gives a very clear, concise and simple account of the particular case of rectification considered, but the treatment is only accurately applicable to very small voltages. The discussion of the grid condenser value assumes a maintained input voltage, and would require some modification in the case of a resonant input voltage.

33. **A Generalised Analysis of the Triode Valve Equivalent Network :** F. M. COLEBROOK : *J. Inst. Elect. Eng.*, 1929, **67**, 157-69.

The paper gives a generalised analysis of the effect of the anode-circuit load on the input impedance of a triode valve. An appendix to the paper discusses the coupling of a tuned circuit to a valve by means of a variable capacity. It is shown that there is an optimum value of this capacity with respect to the potential difference produced at the grid, the optimum condition being such that the effective resistance of the tuned circuit is doubled by the load imposed by the valve. It appears that in many cases this optimum coupling capacity will be less than that appropriate to grid rectification, assuming a maintained input E.M.F., suggesting that in such cases the grid condenser can be reduced without loss of signal strength and with a gain in uniformity of rectification with respect to modulation-frequency output.

(See also Abs. 1.2.2/24.)

2.3.3. TETRODE VALVE RECTIFICATION

1. **The Dynatron :** A. W. HULL, E. F. HENNELLY and F. R. ELDER : *Proc. Inst. Radio Eng.*, 1922, **10**, 320-43.

The dynatron is a three-electrode valve in which the grid is used as the anode and the anode is used merely as a source of secondary emission and is referred to as the dynode. The above paper is actually more concerned with a four-electrode valve, called a pliodynatron, which is similar to the dynatron but has, in addition, an inner or control grid which has the usual function.

As a heterodyne detector, the input circuit is connected to the control grid, without grid condenser or " grid-leak," the anode (i.e., outer grid) to a source of high potential, and the " dynode " through a tuned local oscillator circuit and telephones to a high tension tapping. The local oscillator circuit oscillates through the agency of a negative resistance effect due to secondary emission. An explanation of the action is given, supported by analysis and by oscillographic records. It is shown that the local oscillator amplitude remains constant under the action of the signal.

It is claimed for this arrangement that the filament can be heated by alternating current without any audible disturbance, and that stray disturbances of other than signal frequency are not rectified and remain inaudible. The valve does not, however, appear to have come into general use.

(See also Abs. 4.3/5.)

2. **A Combined Kenotron Rectifier and Pliotron Receiver capable of operation by Alternating Current Power :** A. W. HULL : *Proc. Inst. Radio Eng.*, 1923, **11**, 89-96.

This is a qualitative description of a tetrode valve in which an oxide coated cylindrical electrode surrounding a filament functions as the indirectly heated kathode of a pliotron, the remaining elements of which are contained in the same glass bulb. The filament and surrounding cylinder act as a rectifier for supplying the high tension for the operation of the pliotron. The latter acts as a receiver in the manner described in the preceding abstract. It is claimed that even with three valves the supply frequency hum is not objectionable.

The arrangement does not seem to have come into general use.

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3. **The Four-electrode Valve :** H. NOZIÈRES and P. GIROUD : *Onde Élect.*, 1924, **3**, 583-90.

The paper gives a qualitative account of the double-grid valve, emphasising the function of the interior grid as a means of reducing the space charge effect and thus reducing the high tension required for normal operation. Typical characteristic curves and other data are given. The rectification process is mentioned but not discussed in detail. A " homodyne " type of circuit is described.

(See also Abs. 1.1.8/2.)

4. **The Marconi Four-electrode Valve and its Circuit :** H. DE A. DONISTHORPE : *Proc. Inst. Radio Eng.*, 1924, **12**, 411-21.

A description of some early types of four-electrode valve is given, but the paper is mainly concerned with the Marconi double-grid valve, and a reflex circuit, stated to be in general use by ships, is described. The theory of operation is described in general terms, but details of characteristics are not given.

5. **A Four-electrode Valve Receiver :** G. L. MORROW : *Experimental Wireless*, 1924, **1**, 520-4.

This paper describes a receiving set in which a four-electrode valve is supposed to act as a high-frequency amplifier and detector. The input circuit is connected direct to the inner grid, the outer grid through the secondary of a radio-frequency transformer to the positive high tension terminal, and the anode through the primary and telephones to the filament.

The description of the operation is qualitative and not very detailed.

6. **Supersonic-heterodyne Receiver, employing a Four-electrode Valve :** A. L. WILLIAMS : *Experimental Wireless*, 1924, **1**, 525-6.

The first valve of this set is a triode with normal connections having a tuned anode circuit. The latter is connected directly to the outer grid of a four-electrode valve. The inner grid of the latter is connected through the local oscillator to the filament. The anode circuit is connected through the local oscillator retro-action winding, and a tuned intermediate-frequency transformer to the positive high tension terminal.

No detailed account of the operation is given.

7. **Negative Resistance Phenomena in Double-grid Valves :** P. AMYE : *Onde Élect.*, 1925, **4**, 297-305.

The paper refers to four-electrode valves designed specifically for the production of negative resistance characteristics (e.g., pliodynatron) and then describes the means of obtaining similar results with ordinary double-grid valves. A heterodyne receiving arrangement based on this principle is described qualitatively. It would appear, however, that the ordinary double-grid valve is not particularly suitable for negative resistance operation.

(See also Abs. 1.1.8/3.)

8. **Four-electrode Receiving Valves :** A. C. BARTLETT : *Wireless Wld.*, 1925, **17**, 773-4.

This is a very useful descriptive account of the fundamental principles of the operation of the double-grid valve. Characteristic data are given, but the application to rectification is not described in detail, beyond the facts that low anode voltage only is required and that low resistance loads should be employed.

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9. **The Valve Reception of Modulated Waves without Retro-action :** C. KUHLMANN : *Jahrb. drahtl. Telegr.*, 1925, **25**, 70-82.

This paper is concerned entirely with the grid-circuit rectification of a modulated continuous wave, assuming an exponential form for the characteristic. The performance is analysed, and the conclusions illustrated by measurements carried out are described in Section IV. It is shown that for any given modulation frequency there will be a well defined optimum value of the grid condenser, this optimum value depending on the modulation frequency and the carrier wave frequency. This obviously implies some degree of frequency distortion in the grid rectification of telephony modulated waves.

It is further shown that the modulation frequency output is very approximately proportional to the modulation percentage.

This is a very thorough paper, and gives useful analysis and experimental data on its particular subject. The conclusions, however, refer to a given input E.M.F. and would not necessarily be applicable to the rectification of a resonance potential difference.

(See also Abs. 4.3/11.)

10. **The Valve Reception of Modulated Waves, with Retro-action :** C. KUHLMANN : *Jahrb. drahtl. Telegr.*, 1925, **25**, 95-109.

This paper is concerned exclusively with the grid rectification of modulated continuous waves. A number of conclusions are arrived at analytically and illustrated experimentally, but unfortunately they refer to a rather extreme case (6,000 metre carrier wave, 7,500 cycle modulation frequency), and the relevance of the conclusions to more normal conditions of operation is not discussed. It is shown that retro-action under the given conditions will not in itself increase the output signal strength, since the reduction in effective damping in the input circuit is selective in action and will only increase the carrier wave amplitude, and not the amplitudes of the side-band waves. The measured results show, however, an appreciable increase in output signal strength with increased retro-action, but this is attributed to the establishment of more favourable rectification conditions through the agency of the larger carrier amplitude. Experimental confirmation of a qualitative character is offered in support of this view.

The paper suffers somewhat from the complexity of the problem considered and from the specialisation of chosen carrier and modulation frequencies. It is a useful and suggestive contribution to the subject, but the treatment cannot be considered as entirely conclusive or exhausting. The problem would be more suitably investigated in a simpler form.

(See also Abs. 4.3/9.)

11. **Broadcast Reception with Four-electrode Valves :** F. L. DEVEREUX : *Wireless Wld.*, 1926, **18**, 512-6.

12. **Four-electrode Portable Receiver :** F. L. DEVEREUX : *Wireless Wld.*, 1926, **18**, 544-8.

These are constructional articles describing complete receiving sets using double-grid valves and operating in the ordinary manner but with low anode battery voltage.

No detailed analysis of operation is given.

13. **The Double-grid Modulator :** M. J. DE MARE : *Onde Élect.*, 1926, **5**, 150-61.

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14. **The Double-grid Frequency Changer :** R. BARTHÉLEMY : *Onde Élect.*, 1926, **5**, 162-68.

15. **Note on Modulation in Receiving Apparatus :** H. DE BELLESCIZE : *Onde Élect.*, 1926, **5**, 168-72.

16. **Super-heterodyne Frequency Changers :** L. LÉVY : *Onde Élect.*, 1926, **5**, 173-80.

These papers are all concerned with an application of the double-grid valve to supersonic-heterodyne reception, the local oscillator frequency being self generated by the double-grid valve and applied to the interior grid, and the signal being applied to the exterior grid. De Mare endeavours to show that the operation is quite different in character from the ordinary supersonic-heterodyne process, and is supported by BARTHÉLEMY and DE BELLESCIZE. LÉVY, on the other hand, maintains that no new principle is involved. The conclusion of the controversy would appear to be that though there is a difference in detail as between the double-grid and triode operation, the fundamental principle is not really different. In the one case a term $e_1 e_2$ gives rise to a beat frequency equal to the difference in frequency between e_1 and e_2, and in the other a term $(e_1 + e_2)_2$ gives rise to the beat frequency.

Apart from this controversial aspect of the matter, the papers contain useful information as to the application of double-grid valves to frequency changing purposes.

(See Abs. 1.2.4/11.)

17. **Valve Frequency Changers :** R. BARTHÉLEMY : *Rev. Gén. Élect.*, 1926, **19**, 663-70.

In the first part of the paper the author deals with the theory of the frequency changer as used in the usual supersonic-heterodyne circuit in which the local oscillations are applied to the anode circuit of a triode, the grid circuit of which is supplied with the signal oscillations. The second part of the paper deals with the case in which a double-grid tetrode is used. In this case, the outer grid is connected to the aerial circuit and thus receives the incoming signals. The inner grid contains in its connections an oscillatory circuit so coupled to a suitable coil in the anode circuit as to generate the necessary local oscillations. There is also in the anode circuit the usual circuit tuned to the intermediate or beat frequency. Applying the theory developed in the first part of the paper, the author shows that the mechanism by which the intermediate frequency is produced is essentially different from that of the usual supersonic-heterodyne principle. In particular, it is shown that the amplitude of the beats is independent of the beating frequencies, and that the power available in the intermediate-frequency circuit is limited only by the power of the local oscillator.

(See also Abs. 1.2.4/12.)

18. **A Four-electrode Valve Supersonic Circuit :** *Experimental Wireless*, 1926, **3**, 650.

This describes a patented frequency changing circuit using two double-grid valves. In one valve, the local oscillations are generated by what appears to be a negative resistance effect, a single tuned circuit being connected between the anode and inner grid. The local oscillations are impressed on the outer grid of the other valve, the inner grid of this being connected to the signal input tuned circuit through a grid-leak and condenser. A tuned intermediate-frequency circuit is included in the anode circuit of this valve.

There is no detailed description of the operation.

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19. **Employment of Four-electrode Valves :** B. DECAUX : *Onde Élect.*, 1927, **6**, 1-18.

This is a very excellent descriptive article dealing comprehensively with the subject of four-electrode valves of various types. Particular attention is given to the double-grid valve in which the interior grid serves the purpose of reducing the space-charge effect in the valve, and attention is directed to the very low internal resistance of such valves (6,000 ohms or less). Circuits appropriate to plain and retro-active detection are described.

Typical characteristics are given, but the paper does not contain any detailed analysis of the rectification process.

(See also Abs. 1.2.8/5.)

20. **Detection with the Four-electrode Tube :** J. R. NELSON : *Proc. Inst. Radio Eng.*, 1928, **16**, 822-39.

A very complete analysis on conventional lines of the rectifying behaviour of a four-electrode valve with a modulation-frequency impedance in the anode circuit is given, and is illustrated by reference to the characteristics of the particular screen-grid valve considered (an American pattern). Two experimental values for a 10 per cent. modulation of an unstated modulation frequency are given in confirmation of the analysis, and it is stated that such a screen-grid valve,, with about half a megohm in the anode circuit, is satisfactory for rectification, since it can utilise input voltages of several volts without overloading and is capable of supplying 20 to 30 volts to the grid of the power valve.

The analysis given appears to be very complete in form, but the discussion of its practical applications is less so. No detailed performance curves are given, and the variation of performance with the valve and circuit conditions and with the modulation frequency and modulation depth is not readily deducible from the analysis. It is stated that a high anode circuit resistance load is required (about half a megohm or so), and this would appear to incur the possibility of some degree of performance variation with modulation frequency, but nothing is said on this matter.

21. **Screened-grid Valve as a Detector :** A. P. CASTELLAIN : *Wireless Wld.*, 1929, **23**, 492-4.

The continuous wave rectification performance of a screen-grid valve with a resistance load in the anode circuit was measured under various conditions of anode voltage, screen voltage and resistance load. The output curves obtained are too irregular in shape to permit of any simple description, but show that, in general, the output will vary somewhat critically with all the variables mentioned, particularly with resistance.

The investigation is not complete, as no data are given of performance with a modulated input. In so far as it is possible to draw any conclusion, it is that the screen-grid valve with a resistance load is not very well adapted to rectification either in respect of sensitivity or rectilinearity.

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SECTION III—AUDIO-FREQUENCY AMPLIFIERS

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3.1. CIRCUIT ARRANGEMENTS AND GENERAL PROPERTIES

AUDIO-FREQUENCY amplifiers can be divided into two classes :—

(*a*) Voltage amplifiers.
(*b*) Power amplifiers.

The former of these two constitute the bulk of the amplifiers in use, and their function is to amplify without distortion of wave-form. It is important to distinguish between the amplifier and the power stage, the latter being a converting mechanism whose input is electrical potential and whose output is given in terms of watts delivered to the output mechanism.

The three main types of audio-frequency amplifiers are :—

(*a*) Resistance-Capacity-Coupled Amplifiers.
(*b*) Choke-Capacity-Coupled Amplifiers ; and
(*c*) Transformer-Coupled Amplifiers.

3.1.1. RESISTANCE-CAPACITY-COUPLED AUDIO-FREQUENCY AMPLIFIERS

Although the resistance-capacity amplifier is the simplest type from an elementary theoretical point of view, its design and construction present considerable practical difficulties. To obtain an amplification per stage approaching the voltage factor of the valve used, it is necessary to use an inserted anode resistance considerably larger than the internal resistance of the valve, and this means that the high tension battery voltage must be large in order that the normal direct current voltage drop across the anode circuit resistance shall be such that the anode potential is sufficient to operate the valve at its efficient characteristic point. The use of low factor valves taking anode currents of several milli-amperes limited the use of resistance amplifiers. This limitation has now been removed by the design and use of high factor high resistance

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valves in which the anode current is only of the order of 100 micro-amperes, and in which successful operation is possible with a low anode potential. By this means the high tension battery voltage required has been reduced to quite reasonable values of the order of 100 volts.

Serious progress in the design of resistance-capacity amplifiers commenced when anode resistances as high as three megohms were used (see Abs. 3.1.1/3) in conjunction with valves having voltage factors as high as 35. Since then, progress in design has been rapid, and the high amplification resistance amplifier has rivalled the transformer type.

A resistive anode-circuit load is obviously preferable to that of a choke or transformer due to its frequency independence, assuming that the shunting effect of the following stages is negligible. Owing to the lack of step-up effect, the valve voltage factor must be high if the overall stage amplification factor is to be sufficient to compete favourably with the transformer type. Since this high factor can as yet only be obtained by a valve with a high internal resistance, the inserted anode resistance must be large and the anode current small. The design of the anode resistance has presented problems in itself—the old compound and carbon resistances having a small permissible power dissipation with microphonic difficulties. Wire wound resistances were found preferable, but the self capacity was large and the dimensions such that the component was bulky. The vacuum type of resistance using a metallised filament has to a large extent replaced previous forms of anode resistance owing to its small self-capacity and bulk.

A high factor valve with a large anode resistance gives a large input impedance, and the shunting effect of the complete stage is serious, thereby reducing the effective anode impedance of the preceding stage at high audio-frequencies This effect plays a very important part in the operation of the amplifier and somewhat detracts from the virtues of the high factor valve. The coupling condenser between stages must be large if the low audio-frequencies are to be retained, and this condenser must have a very high insulation resistance. If this condenser is too large, the time constant of the condenser and grid leak circuit may be so large that the system may suffer from temporary paralysis of action due to grid condenser charges obtained by occasional abnormally large signals.

These defects have led to the use of more stages of lower voltage factor valves with lower anode resistances and leaks, the lower anode resistances preventing such serious frequency dependence and the lower leak values tending to maintain stability.

In general :—

(1) If high factor valves are used, a small anode current and high anode resistance is required if the high tension battery voltage is to be kept within reasonable limits. This means that the input impedance of the system will

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be low due to a large stage factor, with consequent reduction in the amplification at high audio-frequencies, i.e., frequency distortion.

(2) If low factor valves are used, the frequency-response characteristic per stage is improved, but a larger coupling condenser is required to maintain the low audio-frequencies. The difficulty here lies in the possibility of grid charges temporarily paralysing the normal action and the use of a greater number of stages. Apart from the economical question of the latter requirement, this reduces the overall amplification at the high audio-frequencies, and it is doubtful whether the net result is much better than with the fewer high factor valves.

The Loewe multiple unit valve (see Abs. 3.1.1/4) represents an attempt to reduce stray wiring capacities to a minimum, but the proximity of the stages introduces retro-active effects and the resistances used are so large, viz., 3 megohms for the anode-circuit resistance and 5 megohms for the grid leak, that the amplification-frequency characteristic is rather unsatisfactory. Portability and simplicity of external connections are the chief merits of the arrangement.

Notwithstanding the difficulties outlined above, the resistance-capacity-coupled amplifier offers a better general amplification-frequency characteristic than the transformer-coupled type, though usually at the sacrifice of amplification. The combination of a resistance-capacity stage which maintains the low audio-frequencies with a transformer stage which can be designed to maintain the higher audio-frequencies appears to give the best arrangement for a two-stage amplifier. A single stage of amplification which can combine the advantages of resistance-capacity coupling with those of the transformer appears to be a desirable solution of the problem of uniform amplification at all frequencies. By such an arrangement both low and high audio-frequency cut-off would be prevented.

3.1.2. Transformer and Choke-Capacity-Coupled Audio-Frequency Amplifiers

The use of a transformer gives the advantage of a step-up effect due to the ratio of secondary to primary turns, but as the ratio is increased and an attempt is made to increase the overall amplification per stage, it is found that the frequency response characteristic becomes peaky. A large value of primary inductance is required if the low audio-frequencies are to be amplified satisfactorily. By suitably adjusting the constants of the transformer, it is possible so to arrange the second resonance point that the higher audio-frequencies are maintained, an advantage which cannot be obtained with a resistance-capacity amplifier.

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The iron cored choke-capacity-coupled amplifier suffers from the defects of the transformer and also eliminates the possibility of any step-up effect; the reduction of the direct current potential drop in the anode-circuit load due to a choke of reasonably low resistance is obtained, and consequently a lower high tension supply will be needed than with a resistive load. Chokes having inductance values up to 300 henries have been advocated in conjunction with high-factor valves, but the difficulties here lie in the serious shunting effect due to the following stage input impedance, which impedance may produce resonance effects and will materially affect the linearity of the frequency-response characteristic. A combination consisting of two chokes and a coupling condenser has been adopted called the "Hiler" circuit. The one choke and the condenser form the usual coupling arrangement, the other choke taking the place of the grid leak of the following valve. By suitably adjusting the values of the three components, it is claimed that the amplification-frequency characteristic can be materially improved and made more nearly linear.

3.1.3. Tuned Audio-Frequency Transformer-Coupled Amplifiers

The tuned audio-frequency choke or transformer-coupled amplifier is only used for the reception of continuous wave signals where great selectivity is required, and usually where Morse recording mechanism is inserted in the output stage. The heterodyned signal is passed to a rectifier and the beat note is applied to the selective audio-frequency amplifier where the "wanted" beat note is amplified at the natural frequency of the tuned system far more than any other beat notes. Such arrangements are in general use for the final selection of weak continuous wave stations operating a recording mechanism. The transformers coupling the valves may be tuned, or filter circuits consisting of complex impedance networks (see 1.1.9) may be used, in which case the frequency selection may be carried even further.

3.1.5. The Output Stage

The "power" or "output" stage is now considered as a converting mechanism in which electrical potential is applied and a power output is obtained. The valve and its attendant circuit behave as a relay in which the input potential controls the power output, or, in some cases, so varies the anode current that recording devices are operated by the change in grid voltage. The desire to obtain a constant ratio of anode-grid potential at all audio-frequencies has led to the use of output loads having an impedance far greater than that of the valve; high impedance loud speakers are used in conjunction with low internal-resistance valves. As

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larger acoustic outputs have been demanded, larger electrical power output has been sought, and power valves having high mutual conductances with low voltage-factors have been produced. Since the anode current is high and the output device has a large inductance and consequently consists of a large number of turns of fine gauge wire, devices to by-pass the normal direct component of the anode current are advisable: choke fed anodes or output transformers may be used. The desire for large grid swings over linear portions of the characteristics has led to the development of "push-pull" systems in which valves are operated back to back. This system can handle twice the grid voltage that could be applied to the same two valves connected in parallel, but the valves should be matched and the grid bias carefully adjusted, otherwise serious wave form distortion can occur.

The tendency has been towards making the impedance of the power output mechanism very large compared to that of the valve, but such a tendency is based only upon empirical assumptions. In the absence of any accurate information regarding the acoustic output-electrical input response characteristic of a loud-speaker, this assumption is justifiable, but when such information is obtained, it may be found that the present criteria for optimum conditions are untenable and both the power stage and the amplifier will need to be designed to compensate for the imperfections of the loud speaker.

3.1.6. Audio-Frequency Amplifiers using Tetrodes

It has been shown that the screen-grid tetrode as at present available is unsuitable for audio-frequency amplification owing to the peaked frequency-response characteristic obtained when used with iron-cored transformers, due to the high internal resistance of the valve, and due to the practical difficulty of using a very large high tension voltage if resistance-capacity-coupling is used. A large anode resistance comparable with the high valve resistance is required with a consequent large direct current anode resistance drop. The negative portion of the characteristic produced by the secondary emission from the anode may be used to give very high amplification ot the order of 150 per stage, but this condition is critically dependent upon adjustments, and only a few hundredths of a volt can be applied; it is therefore of little practical utility.

The pentode possesses the advantage of incorporating an amplifying stage and an output valve in one bulb, owing to its high voltage factor and high mutual conductance. The kink in the characteristic obtained with a tetrode due to secondary emission is eliminated by an earthed screen situated between the anode and its screen. The high voltage factor gives a large input impedance and great care must be taken with the arrangement of this valve if the amplification at high audio-frequencies is to be maintained.

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3.1.7 Filter Circuits.

Filter circuits are used in audio-frequency amplifiers to produce great selective properties, e.g., when an audio-frequency amplifier is used to receive a particular audio-frequency only, such as that obtained from a supersonic-heterodyne receiver. In this case the filter circuit increases the overall selectivity of the receiver. The design of such filter circuits is similar to the design for radio-frequencies (see 1.1.9).

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3.1.1. Resistance-Capacity-Coupled Audio-Frequency Amplifiers

1. **Resistance-capacity Amplification :** B. L. Stephenson : *Wireless Wld.*, 1922, **10**, 18-19, 46-47, Disc. 37-42, 74-77.

This paper briefly discusses from a practical standpoint the use of inter-valve couplings made up of a high resistance and a capacity. The paper concludes with a description of a two-valve amplifier with a modified resistance-capacity coupling. Details are given of the values of the resistances and capacities employed.

2. **Concerning different Types of Amplifiers :** L. Brillouin : *Radio électricité*, 1923, **4**, 109-13.

The various types of valve amplifiers using various coupling arrangements are roughly considered and their relative advantages and disadvantages are briefly discussed.

3. **Resistance-coupled Amplifiers :** M. von Ardenne and H. Heinert : *Jahrb. drahtl. Telegr.*, 1925, **26**, 52-54.

The general principles underlying the design of resistance-capacity-coupled amplifiers are considered, and illustrated by reference to certain German valves and components. A diagram is given showing the use of double-grid valves in a resistance-capacity amplifier. The claim is made that, by a suitable choice of valves and the correct values of the components, a degree of amplification can be obtained equivalent to that with transformer coupling.

High impedance valves with a voltage factor of 35 were used, the anode resistance, coupling condenser and leak being 3 megohms, 0·0005 microfarad, and 10 megohms respectively.

This paper represents an important contribution to the progress in high factor, high impedance valve design, but it was not then appreciated that the shunting effects of the valves would be so serious as they are. No amplification-frequency characteristics are given, and it is clear that M. von Ardenne did not appreciate the more difficult problems associated with this type of amplifier, the disadvantages of which have now been more fully investigated.

4. **The Loewe Resistance-capacity Amplification Unit :** S. Loewe and M. von Ardenne : *Jahrb. drahtl. Telegr.*, 1926, **27**, 19-20.

A general description is given of the rectifier and audio-frequency stages using two double-grid valves in one unit with anode resistances of 3 megohms, coupling condensers of 300 $\mu\mu$F. and grid leaks of 5 megohms,

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together with the circuit arrangements necessary for connecting to the resistance-capacity aperiodic radio-frequency unit.

The frequency-response characteristic of this amplifier element compares unfavourably with that of a well designed resistance-capacity audio-frequency amplifier using separate components.

(See also Abs. 1.1.6/3.)

5. **Resistance-coupled Amplifiers for Audio-frequency Oscillations :** A. FORSTMANN : *Jahrb. drahtl. Telegr.*, 1926, **28**, 156-61.

A discussion is given of the design of an amplifier with resistance-capacity coupling in which the frequency distortion over an assigned audio-frequency range shall not exceed a given amount.

6. **High-note Loss in Resistance Amplifiers :** A. L. M. SOWERBY : *Wireless Wld.*, 1928, **22**, 213-6, 251-3, 285-8.

(See Abs. 3.2.2/23.)

7. **Electrical Characteristics of Amplifiers for Broadcasting Transmitters and their relation to Acoustic Quality :** H. RUKOP : *Jahrb. drahtl. Telegr.*, 1928, **32**, 18-22, 65-70, 86-93.

(See Abs. 3.2.2/27.)

3.1.2. TRANSFORMER AND CHOKE-CAPACITY-COUPLED AUDIO-FREQUENCY AMPLIFIERS

1. **Valve Amplifiers :** M. LATOUR : *Bull. Soc. franç. Élect.*, 1919, **9**, 465-84.

This paper presents a general outline of the construction of various types of amplifiers, and also gives a simple theory of operation, particular attention being given to the best operating conditions for different purposes.

2. **Concerning different Types of Amplifiers :** L. BRILLOUIN : *Radio électricité*, 1923, **4**, 109-13.

(See Abs. 3.1.1/2.)

3. **Distortion in Amplifiers :** G. W. SUTTON : *Wireless Wld.*, 1925, **17**, 156-9.

This paper is chiefly concerned with the performance of various types of inter-valve coupling and their behaviour at various frequencies. Tests were carried out to compare a 33 henry choke with a 70,000 ohm wire-wound resistance when used with a general purpose dull-emitter valve, and further tests were carried out on transformers of various makes. The results are given graphically and discussed. It is shown that the primary inductance of a transformer, as well as the self-capacity, leakage, and ratio of turns, are all important factors which affect the performance of the transformer. The paper concludes with a brief account of the method of measurement employed in the tests.

The superiority of the 70,000 ohm resistance from a frequency-distortion point of view is well brought out by the curves given.

(See also Abs. 4.4.1/5.)

4. **Valve Characteristics :** E. GREEN : *Experimental Wireless*, 1926, 3, 402-6, 469-76.

(See Abs. 3.2.1/6.)

5. **Speech Amplifier Design :** N. W. MCLACHLAN : *Wireless Wld.*, 1926, **18**, 44-48, 95-98, 137-42, 741-46.

(See Abs. 3.2.3/7, 3.3/12, 3.4/7.)

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3.1.3. Tuned Audio-Frequency Transformer-Coupled Amplifiers

1. **Distortion in Radio-telephony :** E. K. Sandeman and N. Kipping : *Experimental Wireless*, 1925, **2**, 756-64, 811-19.

(See Abs. 4.5/6.)

2. **Peaked Audio-amplifier :** A. Kruse : *Q.S.T., Franç. Radioélect.*, 1926, **6**, 29-32.

An account of the development of an audio-frequency amplifier having a peaked amplification-frequency characteristic, suitable for amateur short-wave continuous-wave work.

3. **A Selective Low-frequency Amplifier :** A. Pagès : *Onde Élect.*, 1926, **5**, 276-83.

An amplifier is described in which very sharp and easily adjustable resonance effects may be obtained. This is done by the use of retro-action, such being obtained by means of a transformer whose primary winding is in the anode circuit and whose secondary winding is again transformer-coupled to the grid. Across the primary or anode side of this retro-action transformer is connected a system of impedances which, in the particular case considered, is a simple tuned circuit consisting of an inductance and a condenser in series. It is the frequency to which this circuit is tuned which determines the selective frequency of the amplifier. The input circuit is as follows : there is an input transformer whose primary is connected to the sources of the currents to be amplified—for example, in the case of radio signals it would be connected in the anode circuit of the rectifier ; across the secondary of this input transformer is a resistance ; tapped off across a portion of this resistance are the windings of two other transformers, these windings being in series with one another. The secondary of the first of these windings is connected to the grid and filament ; the other winding referred to is the secondary of the retro-action transformer previously mentioned. The output to the next valve or to the telephones is by means of still another transformer whose primary is in series with the anode, the retro-action transformer and the high tension battery.

The author claims that, due to the method of placing the tuned circuit across the primary of the retro-action transformer, a step-up effect is produced which allows the use of quite small values of inductance and capacity, thus permitting the use of low resistance inductances with their consequent advantages.

4. **Valve Characteristics :** E. Green : *Experimental Wireless*, 1926, **3**, 402-6, 469-76.

(See Abs. 3.2.1/6.)

3.1.4. Special Types of Audio-Frequency Amplifiers using Triodes

1. **Amplifiers for Direct Current and for Current of very Low Frequency :** H. Abraham and E. Block : *C.R. Acad. Sci. Paris*, 1919, **168**, 1321-3.

The paper describes a direct current or very low audio-frequency amplifier with battery coupling between stages.

The anode circuit resistances used had a value of 50,000 ohms and it was found necessary to shunt these with condensers having a capacity of about 0·25 μF. in order to avoid retro-action from the very low-frequency amplifier on the valve preceding it.

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2. **The Kallirotron :** L. B. TURNER : *Radio Rev.*, 1920, **1**, 317-29.

The arrangement here described consists of a combination of circuits, which may be aperiodic, including two ordinary triodes in which secondary emission plays no part, whereby electrical resistance is annihilated—the result achieved in the dynatron by dependence on secondary emission. The theory of the arrangement is first given and then means of obtaining stability are discussed. The instrument, photographs of which are given, may be used as an audio-frequency amplifier, as a telephone amplifier, as a limiter amplifier, and also as an oscillator. Diagrams are given for the various cases.

3. **Amplifiers at very Low Frequency :** P. LEJAY : *Onde Élect.*, 1923, **2**, 521-34.

A scheme for an audio-frequency amplifier for Morse reception is described, and the author states that it is necessary to study two phenomena in this circuit, since :—

(1) At the commencement of signalling, the apparatus only functions after the lapse of a certain definite time, the value of which depends upon the values of capacities and resistances in the various circuits. This lag may be of several seconds duration.

(2) When signalling has ceased, or during intervals between words, one, two or three parasitic records may occur which are certainly not due to natural atmospherics. To investigate these effects, an apparatus was constructed having one detecting valve and four amplifying valves, the circuit constants being given. An oscillograph not being available, records of anode current were obtained from a milliammeter, and many tracings of these records showing the Eiffel Tower signals are given. The adjustment of the circuit constants to give an instrument free from these defects is then described and records are shown. A theoretical treatment of the action of capacities in shunt to the anode circuit resistance is then given.

4. **Applications of a New Form of Amplifier :** P. LEJAY : *C.R. Acad. Sci. Paris*, 1927, **185**, 500-2.

The device described by the author is similar to the Kallirotron in that it consists of two valves so arranged that the anode resistance of the second one forms part of the grid circuit of the first one. The first valve is resistance-capacity coupled to the second one. By suitably choosing the resistances, it is possible to use the combination for detection, oscillation, amplification and as a limiting device.

5. **Distortionless Audio-frequency Resistance Amplifiers :** H. F. MAYER : *Elektrotech. Z.*, 1927, **48**, 10-12.

(See Abs. 3.2.2/17.)

6. **Low Frequency Amplification with Choke Coupling :** H. KAFKA : *Jahrb. drahtl. Telegr.*, 1928, **31**, 87–90.

(See Abs. 3.2.3/14.)

7. **Some Improvements in Direct-current Amplifiers :** R. JOUAUST and B. DECAUX : *Onde Élect.*, 1928, **7**, 306-8.

The overall amplification due to a valve and a resistance-coupling unit is given by the expression $K.R./(R \times \rho)$, where K is the amplification factor of the valve, ρ its internal resistance and R the resistance of the anode circuit. This expression reaches a maximum value K when R is indefinitely high, that is, when the anode current through the valve is constant. This condition is realised by using a saturated diode as the anode resistance.

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By using a special triode and a particularly suitable diode, voltage amplifications of the order of 200 to 300, as measured on an electrometer, were obtained.

3.1.5. The Output Stage

1. **Amplifiers of the Push-pull Type :** H. Chireix : *Radioélectricité*, 1923, **4**, 65-68.

A short article with illustrations dealing with the circuit used with the balanced carbon microphone as employed by the Western Electric Company. Details are given of push-pull amplifier circuits with both transformer and resistance-capacity coupling. Mention is also made of the balanced modulator.

2. **Push-pull Speech Amplifiers :** W. James : *Wireless Wld.*, 1924, **14**, 271-7.

An article describing in detail how to construct an audio-frequency amplifier working on the push-pull principle, and describing the advantages of such an arrangement in minimising frequency distortion.

3. **A New Method of Push-pull :** L. E. T. Branch : *Wireless Wld.*, 1928, **22**, 547-50.

It is shown that the customary output transformer may be dispensed with in the push-pull circuit if we have a centre-tapped loud speaker winding. A brief numerical analysis is given in which it is shown that the powers radiated by a loud speaker when driven by one valve, two valves in parallel, or two valves in push-pull, are in the ratios of 1 : 1·5 : 3.

4. **The Output Stage and the Pentode :** N. W. McLachlan : *Wireless Wld.*, 1928, **23**, 30-33, 77-80, 113-6.

(See Abs. 3.2.5/19, 3.2.6/6.)

(See also Section 3.2.5.)

3.1.6. Audio-Frequency Amplifiers using Tetrodes

1. **The Resistance-coupled Amplifier :** H. Kröncke : *Wireless Wld.*, 1925, **17**, 395-8.

(See Abs. 3.2.2/4.)

2. **The Screened Valve in Low Frequency Circuits :** N. W. McLachlan : *Wireless Wld.*, 1927, **21**, 536-8, 577-79.

(See Abs. 3.2.7/2.)

3. **The Output Stage and the Pentode :** N. W. McLachlan : *Wireless Wld.*, 1928, **23**, 30-3, 77-80, 113-6.

(See Abs. 3.2.5/19, 3.2.6/6.)

4. **A New Method of Using Resistance Amplification with Screened-grid Valves :** J. J. Dowling : *Experimental Wireless*, 1928, **5**, 61-62.

(See Abs. 3.2.7/3.)

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3.1.7. Filter Circuits

1. **The Design of Filters for Audio-frequencies :** C. A. Beer and G. J. S. Little : *J. Post Office Elect. Eng.*, 1925, **17**, 298-317.

The paper gives a large number of curves of the ratio of the current at the end of a filter to the current at the beginning plotted against the frequency ratio W/W_o, where W_o is the cut-off frequency, for high-pass and low-pass filters, including a mutual inductance filter, for various numbers of links in the filter chain and under various conditions as regards terminal impedance. Some of these curves are theoretical ones, but some are obtained from actual observations on filters constructed with air-core inductors and with iron-core inductors. In the latter case the dimensions of the stalloy stampings found suitable are given ; it is found advisable to include an air-gap in the magnetic circuit. The paper is intended to assist those who wish to use simple filter circuits without going into the mathematics of the subject.

2. **Tone Correction in Low-frequency Amplifiers :** H. Lloyd : *Wireless Wld.*, 1926, **18**, 11-15.

(See Abs. 3.4/10.)

3. **Radio Interference :** R. B. Ashbrook and R. W. Wright : *Elect. Wld.*, 1926, **88**, 851-3.

(See Abs. 1.2.10/8.)

4. **Search for Good Quality in Selective Receivers :** P. Borias : *Bull. Soc. franç. Radio*, 1928, 3-12.

A series of coupled audio-frequency circuits is recommended, a resultant tuning curve with a flat top being obtained by suitable shifts of the tuning of each circuit away from the mid-point.

(See also Section 1.1.9.)

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3.2. ANALYSIS OF AUDIO-FREQUENCY AMPLIFICATION

3.2.1. General Theory and Transients

The transient phenomena occurring in audio-frequency amplifiers is somewhat similar to that occurring in radio-frequency amplifiers (see 1.2.1), but special forms of transient applicable only to audio-frequencies can be considered as types of distortion. Most of the references applicable to such transients are contained in 3.4.

3.2.2. Resistance-Capacity-Coupled Audio-Frequency Amplifiers

The analysis of the resistance-capacity-coupled audio-frequency amplifier prior to 1925 assumed the valve to possess certain definite characteristics which could be represented by the well-known four parameters of a valve. By neglecting all inter-electrode capacity and retro-active effects, simple expressions were obtained which were used to design the anode resistances and coupling condensers. Progress in the design of resistance-capacity-coupled amplifiers was made by M. von Ardenne (see Abs. 3.2.2/3), who in 1925 made use of anode resistances as high as three megohms with valves having a voltage factor of 35. H. Kröncke (see Abs 3.2.3/4) published a summary of this work and pointed out that prior to Ardenne's researches the assumption was made that the anode voltage excursions must lie on the linear portion of the characteristic curve, i.e., that the anode resistances used should be of such a magnitude that the drop in potential in the anode resistance would leave sufficient anode voltage to ensure that the valve would operate in the region of its linear characteristic. It is then shown that this condition is not necessary and that, even if the anode voltage is so low that its characteristic working point falls in the curved portion of the static characteristic, the dynamic characteristic may be linear. Colebrook (see Abs. 3.2.2/19) has given a mathematical analysis of a resistance-capacity stage working under these conditions and has shown that a linear response characteristic can be obtained.

Most of the early German work neglected the effects due to the inter electrode capacities, which effects are now known to play an all-important part in the behaviour of any audio-frequency amplifier, particularly at high audio frequencies. The effect of this input circuit load upon the preceding stage is materially to reduce the amplification, and since this load is largely dependent upon the voltage factor of the valve following the stage under consideration,

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it is now appreciated that the use of high-factor valves is not altogether satisfactory, if a constant amplification-frequency characteristic is required. Much diversity of opinion exists regarding the best values of anode resistance, coupling condenser and grid leak. The use of a high anode resistance gives a stage amplification factor approaching the voltage factor of the valve, but the shunting effect of the following valve is serious, and at high audio-frequencies the amplification is seriously reduced. A low anode resistance gives a more uniform amplification-frequency characteristic at the sacrifice of amplification. Again, the question of the best value of the coupling condenser is one which is much disputed. A small condenser gives a cut-off at low audio-frequencies, whereas a large condenser gives a large time-constant to the grid leak circuit, and should the grid ever become positively charged, due to an unusually large impressed signal, the distorting effect, due to the slow leakage of accumulated charge on this condenser, may be serious. This effect is unlike that which produces a periodic distortion; the distortion is more like that due to a temporary and partial paralysis of the normal operating behaviour. That such temporarily large signals occur, however large the grid bias battery may be, is certain, and it is necessary to consider such abnormal effects in designing an amplifier.

High resistance leaks, although theoretically advantageous, tend to promote instability; thus a compromise must be made, but the optimum values are by no means agreed upon. Very little experimental data is available and there is much need for an investigation into certain subsidiary properties, such as instability, transient phenomena, audio and sub-audio paralysis due to grid charges, valve howling at audio and super-audio-frequencies, microphonic noises, etc. It is impossible at present to obtain a good amplification-frequency characteristic with more than three stages, and the limits set to the possibility of increasing the amplification are due to effects which can probably be eliminated, in which case the limit will be set by the " Schrot " and " Johnson noise " effects only (see 1.2.6).

The advantages to be obtained by screening the individual stages are now well realised. Such screening helps to prevent stray inductive and capacitative couplings between various parts of the amplifier and tends to assist stability by removing causes of regeneration. It is unfortunate that the screen-grid tetrode in its present form is unsuitable for audio-frequency work, except in a selective frequency amplifier, since its small anode-grid capacity would help to reduce retro-active effects.

The investigations of M. VON ARDENNE and HEINERT (see Abs. 3.2.2/3) started from the view that a resistance-capacity amplifier was purely a voltage device in which no power was taken from the input and they drew a distinction between " power " and " voltage " amplifiers. However, since the input impedance of any stage is

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not purely reactive, some power is expended and the "voltage" amplifier modifies the circuit to which it is connected. It is becoming increasingly difficult to regard an amplifier of any type as a unit whose properties are independent of the input and output circuits, and the tendency is rather to express the properties in the form of

(*a*) The voltage amplification with a definite known output circuit; and

(*b*) The input impedance with this known output circuit. Miller's work on input impedance (see Abs. 1.2.7/2) has materially helped to show that the amplifier cannot be regarded purely as a unit, since the nature of the output load, which is electrically coupled through the anode-grid capacity of the power valve to the output circuit of the amplifier, influences the effective impedance of this output circuit. Similarly, the input impedance of the complete amplifier influences the properties of the preceding rectifier and therefore any analysis of the properties of the amplifier must commence with the power valve output load and trace its effect, coupled with the normal impedances of each stage, through the stages to the input circuit.

The first stage of a three valve resistance-capacity amplifier produces the most perceptible frequency distortion due to the cumulative loading effects of the following two stages and the nature of the output load (see Abs. 3.2.2/12). Several authors (see Abs. 3.2.2/12, 18) have commented upon the ambiguity of publishing single-stage characteristics, as these are never maintained when the stage is followed by other stages which introduce loading and regenerative effects.

3.2.3. Transformer and Choke-Capacity-Coupled Audio-Frequency Amplifiers

The advantage to be obtained by the use of transformer coupling viz., a step-up effect, so that the stage amplification factor is greater than the voltage factor of the valve, is obtained only at the expense of certain disadvantages. The frequency-response characteristics of the early types of inter-valve transformers were so poor that resistance-capacity-coupling was vastly superior, but, due to the development of transformers with large primary inductance and low secondary self-capacity, the transformer-coupled stage can now give a characteristic which in some ways is preferable to that of the resistance type. In general, the transformer-coupled stage is found to possess a frequency-response characteristic which has two peaks corresponding to the first and second resonance points of the transformer. The peakiness and frequency at which these points occur are largely dependent upon the internal resistance

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of the valve, and by suitably arranging the valve constants, it is possible to damp out almost completely the first resonance point and maintain the second one at a frequency in the neighbourhood of 4,000 p.p.s. or over. The characteristic is therefore lifted at the upper frequencies where the resistance-capacity stage tends to fall off in amplification.

Although a well-designed resistance-capacity-coupled stage using low-factor low-internal resistance valves gives a better frequency-amplification characteristic than the transformer type, a well designed transformer stage usually gives a better response characteristic than can be obtained with a resistance-capacity-coupled stage using high factor valves owing to the shunting effect of the stray and inter-electrode capacities. The combination of the resistance and transformer types gives the best arrangement for an amplification up to about 500 times the input voltage.

The desire to obtain linear frequency-amplification characteristics from all types of audio-frequency amplifiers is due to a lack of exact knowledge regarding the response characteristic of the loud-speaker and the assumption that the amplifier is perfect if the linear relationship holds. Since the amplifier is a link in the chain, its properties are dictated by the apparatus preceding it, and the loud-speaker which follows it. A good deal is known about the apparatus that lies before the amplifier but little is known about the loud speaker performance, and until exact information is forthcoming, the design of amplifiers will continue to rest upon the assumption that uniform amplification at all audio frequencies means that the amplifier is perfect.

The ultimate quality from a loud speaker depends upon—

(*a*) The response-characteristic connecting the input acoustic energy to the microphone with the modulated energy in the transmitting antenna.

(*b*) The overall frequency-amplification of the receiver.

(*c*) The energy-frequency characteristic of the loud speaker in free space.

(*d*) The "loudness level," which includes the physiological aspect of the matter.

(*e*) The acoustic properties of the room in which the loud speaker is operated.

From a consideration of (*a*) and (*c*), it does not follow that the overall characteristic of the receiver should give equal amplification at all frequencies. Certain types of loud speaker require an amplifier characteristic which dips in the middle audio register to give due prominence to the low and high audio frequencies, the amount of such dip depending on the "loudness level." For other types, the amplifier should give a prominent low audio-frequency resonance to compensate for the insensitivity of the speaker at these frequencies.

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Having obtained the most satisfactory characteristic for a given "loudness level," there are still two salient defects :—

(*a*) The absence of stereophonic effect, or the sense of directivity and location.

(*b*) Resonances introduced at various parts of the system.

A variable overall characteristic can be secured by various artifices inserted in the valve coupling units, such devices taking the form of filter circuits. So far as low-toned instruments are concerned, a good fundamental is required for realism ; they should not be " sensed " by their overtones. Comparatively inconspicuous resonances often become prominent with increasing intensity.

The time required for the resonating component to decay to inaudibility increases with the intensity, and with the diaphragm type of loud-speaker the lower the frequency of the resonance note, the more sustained is its effect, because the damping is usually less for the low than for the high tones ; hence, at large intensities, the time taken for the natural vibration to die away to inaudibility is usually appreciable and causes an unpleasant " blurring " distortion of the output. Horn resonance is not particularly objectionable for weak sounds, but with loud sounds the low tones are usually badly defined. Appreciable response causes a " blurring " of the band of frequencies on each side of the fundamental, and heavy retro-action in the amplifier causes a similar distorting effect. Owing to the lack of exact information upon these important points, there is a need for a thorough examination of the acoustic aspect of the subject with a view to defining the required characteristics of the amplifier.

It should be pointed out, however, that it is only an expedient to correct the faults of the loud speaker by a suitable amplification-frequency characteristic from the amplifier. The correct procedure is to develop a loud speaker possessing the required characteristics, but since this is more difficult than a modification of the audio-frequency amplifier, such modification is justifiable at present.

The difficulty of this method of " compensation " is that a particular amplifier would need to be designed for every different loud speaker.

3.2.4. Audio-Frequency Filter Circuits

The theory of audio-frequency filter circuits has already been dealt with (see 1.1.9).

3.2.5.(*a*) The Power Stage and Power Amplifiers

The power stage is now looked upon as a converting mechanism and, since the output demanded is now so large, valves having a high mutual conductance and a long characteristic straight

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portion are needed. The "Push-Pull" method can be adopted. There are two methods of obtaining a linear characteristic from two valves operated back to back :—

(*a*) With a grid bias which sets both valves almost at their rectifying points.
(*b*) With a grid bias which sets the valves at the mid-points of their linear characteristic parts.

If the curved portion of the valve static characteristic is parabolic, the correct grid bias for (*a*) is half way along the abscissa of the curved portion. The permissible grid swing in (*a*) is twice that obtainable in (*b*). In (*a*) the valves must be matched, but in (*b*) the two characteristics should merely be linear over the same values of grid swing : the slopes of the characteristics need not be identical though it is preferable that they should be matched.

3.2.5.(*b*) The Pentode

A recent development of the tetrode is the pentode, the essential properties of which are those of a tetrode, the valve possessing a high voltage factor, a high impedance, and a high mutual conductance. The kink in the normal characteristic of a tetrode due to secondary emission from the anode is eliminated by an earthed screen situated between the anode and the normal screen, this screen preventing secondary electron emission to the normal screen. It is claimed that the pentode can replace an ordinary amplification stage and the power stage (see Abs. 3.2.5/19), and if the output load is designed to suit the high internal resistance of the valve, better output stage performance can be obtained ; but great care must be taken over input lead connections, since the high factor valve gives a large input impedance which, thrown as a load on the preceding stage, may nullify the advantages obtained by the use of the pentode.

3.2.6. Retro-Action

A large amount of rectro-action in audio-frequency amplifiers is due to the inherent inter-electrode capacities, and the output load is transferred by this coupling to the previous stages, but much of the retro-action experienced is due to common batteries, stray coupling, etc. The common battery resistance reduces the amplification if the number of stages is odd and increases the amplification, tending to produce instability if the number of stages is even (see Abs. 3.2.6/2). These effects are due to the opposite phase relationships existing between the grids and anodes of each stage.

To prevent such effects, methods of anode feed are used (see Abs. 3.2.6/5) in which an impedance of low value shunts the battery circuit, which is deliberately made high by the insertion of series feed

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resistors in each anode circuit. Battery eliminators tend to produce instability due to the common impedance of the eliminator (see Abs. 3.2.6/3), and in nearly every case affect the overall characteristic seriously. It has been suggested that such an eliminator can be utilised to correct the loud speaker response characteristic (see Abs. 3.2.6/4), but such a suggestion is based entirely upon empirical views, and no design based upon theoretical considerations is at present possible.

3.2.7. Amplifiers using Special Valves

The use of double grid tetrodes to reduce the necessary high tension voltage has met with little practical success owing largely to the limited length of the linear portions of the characteristics and to the larger anode current required. There seems to be no merit in reducing the high tension voltage if the total energy demands from the battery are not reduced as well.

McLachlan has shown (see Abs. 3.2.7/2) that there is nothing to be gained by the use of existing types of screen-grid tetrodes at audio frequencies. With resistance-capacity coupling, the high tension battery voltage required is excessive owing to the high internal resistance of the valve, and, with transformer coupling, the amplification-frequency characteristic is very peaked due to the same cause. Although the kink in the characteristic due to secondary emission may be used to obtain an amplification of 150 per stage, the conditions of operation are critical and only a few hundredths of a volt can be applied.

3.2.8. Screening of Audio-Frequency Amplifiers

The problem of screening at audio-frequencies is a more difficult one than that of screening at radio-frequencies, but exactly the same methods are adopted (see 1.2.9.)

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3.2.1. General Theory and Transients

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2. **Effect of Space Charge and Residual Gases on Thermionic Currents in High Vacuum :** I. Langmuir : *Phys. Rev.*, 1913, **2**, 450-86.
(See Abs. 1.2.1/2.)

3. **The Pure Electron Discharge and its Application in Radiotelegraphy and Telephony :** I. Langmuir : *Gen. Elect. Rev.*, 1915, **18**, 327-39 ; *Electrician*, 1915, **75**, 240-5 ; *Elect. Rev. Lond.*, 1915, **76**, 753-5 ; *Lumière Élect.*, 1915, **29**, 241-5, 272-9.
(See Abs. 1.2.1/3.)

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4. **Valve Amplifiers :** F. OLLENDORFF : *Arch. Elektrotech.*, 1924, **13,** 274-91.

(See Abs. 1.2.3/6.)

5. **Faithful Reproduction in Radio-telephony :** L. C. POCOCK : *J. Instn. Elect. Eng.*, 1924, **62,** 791-807.

(See Abs. 3.4/4.)

6. **Valve Characteristics :** E. GREEN : *Experimental Wireless*, 1926, **3,** 402-6, 469-76.

A paper showing how anode-current/anode-voltage characteristics may be used in studying the action of valve circuits by graphical analysis. The first part of the paper describes in detail the method of using such characteristics, and the method is applied to determine the conditions giving maximum output and efficiency. The second part of the paper applies the method to a discussion of (*a*) resistance amplifiers, and (*b*) transformer or tuned-circuit amplifiers, including distortionless power amplifiers. Numerical examples are given, dealing with (i) power valves in parallel and (ii) the modulation on a six kilowatt transmitter using choke control.

The analysis does not take into consideration the effects of inter-electrode capacities.

7. **Equivalent Circuit for Amplifying Valves :** H. F. MAYER : *Telegr. u. Fernspr. Tech.*, 1926, **15,** 335-7.

By analogy with a generator circuit, two simple arrangements are obtained as circuit equivalents for a valve. Application is made to the case of a resistance-capacity coupling between two valves, and the usual expression is obtained for the amplification factor.

8. **Speech Amplifier Design :** N. W. MCLACHLAN : *Wireless Wld.*, 1926, **18,** 44-48, 95-98, 137-42, 741-46.

(See Abs. 3.2.3/7, 3.3./12, 3.4/7.)

9. **Calculation of Anode Current and Amplification Factor in Resistance Amplifiers :** BERMBACK : *Elekrotech. Z.*, 1927, **48,** 757-9.

(See Abs. 3.2.2/16.)

10. **Unstable Equilibrium and Parasitic Phenomena in Triode Circuits :** M. PODLIARSKY : *Onde Élect.*, 1928, **7,** 287-306.

(See Abs. 1.2.1/22.)

3.2.2. RESISTANCE-CAPACITY-COUPLED AUDIO-FREQUENCY AMPLIFIERS

1. **The Kallirotron :** L. B. TURNER : *Radio Rev.*, 1920, **1,** 317-29.

(See Abs. 3.1.4/2.)

2. **Further Notes on Resistance-capacity Amplification :** F. M. COLEBROOK : *Experimental Wireless*, 1924, **1,** 712-6.

The author points out the merits of resistance-capacity amplification over transformer-coupling as regards the constancy of amplification over the whole audible range of frequencies. The necessity for a reasonably large coupling condenser is pointed out, but increasing this condenser above the value required to give a reasonably straight frequency-response characteristic introduces difficulties due to dielectric losses, poor insulation and the time-constants of the circuits.

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3. **Resistance-coupled Amplifiers :** M. VON ARDENNE and H. HEINERT : *Jahrb. drahtl. Telegr.*, 1925, **26**, 52-54.

(See Abs. 3.1.1/3.)

4. **The Resistance-coupled Amplifier :** H. KRÖNCKE : *Wireless Wld.*, 1925, **17**, 395-8.

The general principles underlying the design of resistance-capacity coupled amplifiers are considered and illustrated by reference to certain German valves and components. A diagram is given showing the use of double-grid valves in a resistance-capacity amplifier in which anode resistances as high as 3 megohms can be used. The claim is made that, by a suitable choice of valves and the correct values of the components, a degree of amplification can be obtained equivalent to that with transformer coupling.

No consideration is given, however, to the effects due to inter-electrode capacities, and with such high resistances these effects will be serious.

5. **Valve Characteristics :** E. GREEN : *Experimental Wireless*, 1926, **3**, 402-6, 469-76.

(See Abs. 3.2.1/6.)

6. **Resistance-coupled Amplifiers for Audio-Frequency Oscillations :** A. FORSTMANN : *Jahrb. drahtl. Telegr.*, 1926, **28**, 156-61.

(See Abs. 3.1.1/5.)

7. **The Best Working Conditions for a Resistance Amplifier :** M. MERCIER : *Onde Élect.*, 1926, **5**, 413-24.

An account of an investigation dealing with the various factors affecting the efficiency of resistance-capacity amplification. The results of varying the anode voltage, anode resistance, filament temperature, &c., are shown by means of graphs. The importance of these variations is discussed in a very general manner.

The analysis is based entirely on characteristic curves and all capacity effects are neglected.

(See also Abs. 4.4.1/6.)

8. **Resistance Amplification :** R. HIECKE : *Elektrotech. u. Maschinenb.*, 1926, **44**, 89-92.

The general theory of resistance-capacity amplification is given and the requirements of valves for this type of amplification are stated, together with the part played by the grid condenser and valve capacities. The amplification is proportional to the shape of the anode current-grid voltage characteristic. For high amplification, large values of the grid, anode and internal resistances are recommended. A table is given showing the variation of amplification with frequency for various values of the coupling condenser.

9. **Equivalent Circuit for Amplifying Valves :** H. F. MAYER : *Telegr. u. Fernspr. Tech.*, 1926, **15**, 335-7.

(See Abs. 3.2.1/7.)

10. **Comparison between Transformer and Resistance Amplifiers :** M. VON ARDENNE : *Jahrb. drahtl. Telegr.*, 1926, **27**, 167-9.

(See Abs. 3.2.3/8.)

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11. **Notes on the Design of Resistance-capacity-Coupled Amplifiers:** S. HARRIS: *Proc. Inst. Radio Eng.*, 1926, **14**, 759-63.

An analysis of the coupling in the resistance-capacity-coupled amplifier is given, in which the variation of the voltage ratio with frequency is considered. A method is given for determining the values of the resistances and capacities for which the variation of the voltage ratio over a given frequency range will be a definite and known amount.

12. **On the Values and the Effects of Stray Capacities in Resistance-coupled Amplifiers:** M. VON ARDENNE and W. STOFF: *Proc. Inst. Radio Eng.*, 1927, **15**, 895-901.

The general theory of stray capacities is first developed with due consideration of the phase conditions, and then the values and effects of stray capacities are discussed for a three-stage amplifier of practical dimensions.

The authors show that the overall performance of an amplifier cannot be calculated by considering one stage and multiplying by the number of stages. Each stage must be examined separately, commencing with the anode circuit of the power stage and working backwards through the amplifier to the grid of the first valve. The manner in which the output load affects the input impedance of all stages is examined and it is shown that, with resistance-capacity-coupled audio-frequency amplifiers, it is the frequency-response characteristic of the first stage only that is of material importance.

13. **Design of Audio-frequency Resistance Amplifiers free from Distortion:** M. VON ARDENNE: *Elektrotech. Z.*, 1927, **48**, 36-39.

The paper discusses the design of resistance-capacity-coupled amplifiers for high voltage amplification and efficiency, with special reference to the frequency, valve constants, anode potential, grid potential, &c., high anode resistances of the order of 3 megohms being employed. In the second part, the manner in which the amplification depends upon the frequency and amplitude of the incoming oscillations is investigated, and it is shown that high resistance, high factor valves give no amplitude distortion if the input voltage is small.

14. **A Graphical Method of Amplifier Coupling Design:** M. G. SCROGGIE: *Experimental Wireless*, 1927, **4**, 733-4.

The three main points in the design of resistance-capacity coupling are:—

(i) The drop in amplification due to the impedances of the condenser as compared with that of the grid leak.
(ii) The drop in amplification due to the shunting effect of the leak.
(iii) The time constant of discharge of the condenser.

(i) Is important at the lowest audio-frequencies; (ii) need not be considered at very low frequencies, nor at any frequency provided that the leak resistance is large compared with the anode resistance; (iii) will not enter in if the greatest peak of signal or atmospheric never makes the grid positive.

In practice, a few peaks do occasionally overstep the normal limits and produce momentary grid current. If the charge so produced (on the coupling condenser) can leak off very rapidly, the quality of reproduction will suffer only slightly. Hence this condenser should be as small as possible consistent with (i).

The second condition is open to criticism since the shunting effect of the input impedance upon the leak is neglected, and at high audio-frequencies the effective impedance of the leak may be very materially reduced. Curves illustrating the cut-off to be expected at various audio-frequencies for different values of the coupling condenser are given.

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Since all the factors are not considered, these curves do not represent true conditions. It is stated that it is doubtful whether the human ear can detect less than a 30 per cent. drop in strength of any restricted band of frequencies when mixed with others.

15. **Theory of Audio-frequency Amplification with Resistance Coupling** : H. KAFKA : *Jahrb. drahtl. Telegr.*, 1927, **29**, 39-45.

The frequency dependence of audio-frequency amplification employing resistance coupling may be represented by means of a locus diagram. The valve capacities and the load proportions of the following stage are considered, and the use of the diagram in determining the most suitable resistance combination is indicated.

16. **Calculation of Anode Current and Amplification Factor in Resistance Amplifiers** : BERMBACK : *Elektrotech. Z.*, 1927, **48**, 757-9.

The space charge equation is transformed into a simple cubic and the conditions for maximum amplification and minimum distortion are obtained.

17. **Distortionless Audio-frequency Resistance Amplifiers** : H. F. MAYER : *Elektrotech. Z.*, 1927, **48**, 10-12.

This paper gives a summary of the elementary equations showing the relationship between amplification factor, limits of frequency range and the physical constants of resistance-capacity audio-frequency amplifiers. The conditions for uniform amplification throughout the whole range of audio-frequencies are discussed, and a reactance arrangement for use with resistance amplifiers is described, which increases the amplification factor very appreciably.

No inter-electrode capacity effects are considered.

18. **Resistance-capacity Amplification** : W. JAMES : *Wireless Wld.*, 1927, **21**, 15-20, 43-46, 88-90.

After pointing out that single stage characteristics have little significance, owing to the serious shunt loading effects of the following stage, the author gives the theory of resistance-capacity amplification, and shows that, although the alternating current resistance of the valve increases with the coupling resistance, the voltage amplification is also thereby increased. The effects of the grid leak are to lower the alternating current voltage, which can be applied without distortion, and to reduce the voltage amplification. The method of calculating the capacity of the coupling condenser is given, and the relation between the latter and the grid leak discussed. The paper concludes with a discussion of stray capacities and their effects on the performance of resistance amplifiers.

19. **A New Development in Resistance Amplification** : F. M. COLEBROOK : *Experimental Wireless*, 1927, **4**, 195-205.

The work of M. VON ARDENNE and HEINERT, published by KRÖNCKE, is discussed and the possibility of using very high anode resistances of the order of several megohms with normal high tension battery voltage is suggested and mathematically analysed. It is claimed that distortionless voltage amplification can be obtained when the anode potential is so low that the static characteristics are non-linear. The principle is extended to the case of a multi-stage amplifier and the following conclusions are deduced :—

(*a*) An amplification per stage of from 80 to 95 per cent. of the voltage factor of the valve can be obtained. For input voltages not exceeding 1 volt an amplification of at least 20 per stage can be obtained, using an anode battery voltage of not more than 100 volts.

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(*b*) The amplitude distortion will be small and the frequency distortion will depend on the input impedance of the following stage. To reduce this latter, care must be taken in the lay-out of the wiring and valve-holders so as to reduce capacity effects as much as possible.

20. **Theory of Resistance Amplifiers :** H. DÄNZER : *Jahrb. drahtl. Telegr.*, 1927, **30**, 26-28.

The influence on the amplification ratio of the battery voltages, the ohmic resistance of the anode circuit and the grid leak resistance are investigated. The maximum amplification is not critical, but occurs when $2r_\mu < r_a < 4r_\mu$, in general when $r_a = 2r_\mu$, where r_a is the anode resistance and r_μ the grid leak resistance. The dependence of the maximum value of the amplification ratio on the above factors is also discussed.

21. **Calculation of Apparent Capacity in Resistance Amplifiers :** M. v. ARDENNE and W. STOFF : *Jahrb. drahtl. Telegr.*, 1927, **30**, 86-89.

(See Abs. 1.2.7/8.)

22. **Amplification and High Quality :** N. W. MCLACHLAN : *Wireless Wld.*, 1928, **22**, 463-6.

(See Abs. 3.4/15.)

23. **High-note Loss in Resistance Amplifiers :** A. L. M. SOWERBY : *Wireless Wld.*, 1928, **22**, 213-6, 251-3, 285-8.

The loss of high audio-frequencies is much greater if anode resistances of the order of a megohm are used, together with high impedance valves, than if these resistances are kept fairly low. A formula and curve are given from which the percentage loss of amplification of a signal at a frequency of 5 kilo-cycles per sec. can be calculated for any inter-stage coupling, given certain working values of the circuit. The paper concludes by applying these calculations and data to a circuit consisting of an anode rectifier, a high-amplification valve acting as an audio-frequency amplifier, and a small power valve in the output stage. A circuit designed to avoid serious loss of high audio-frequencies and still retain good overall amplification, is also given.

24. **Calculations for Resistance Amplifiers :** A. L. M. SOWERBY : *Experimental Wireless*, 1928, **5**, 201-3.

An approximate method is given of solving the resistance-capacity-coupled case at audio-frequencies. The solution can be presented in a graphical form and is only a rough approximation for practical work.

25. **A Graphical Construction for Resistance Amplifiers :** W. A. BARCLAY : *Experimental Wireless*, 1928, **5**, 499-500.

The author gives a graphical method of solving the equations deduced by SOWERBY (see previous Abs.), from which the best values of anode resistance, grid bias, external resistance, &c., to secure maximum output for a particular amplifying valve may be calculated.

The article represents an extension of Sowerby's method.

26. **Inter-electrode Capacities and Resistance Amplification :** L. HARTSHORN : *Experimental Wireless*, 1928, **5**, 419-30.

If an audio-frequency resistance-coupled amplifier is required to give distortionless amplification, the components should satisfy the following conditions :—(1) The condenser coupling the anode circuit of one valve to the grid of the next should be of large capacity, say, 0·002 μF, and high insulation resistance ; (2) the anode resistances should be considerably

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larger than the internal resistances of the valves, but should not be of higher resistance than is necessary, and preferably inductively wound to compensate for the input capacity of the following valve; (3) the resistance of the grid leak should be large compared with that of the preceding valve, and should not have a large self-capacity.

The bridge method of measuring the input capacity of a valve system is described in detail and typical values are given for modern valves.

The analysis is very thorough and no factors are neglected or approximated.

27. **Electrical Characteristics of Amplifiers for Broadcasting Transmitters and their Relation to Acoustic Quality:** H. RUKOP: *Jahrb. drahtl. Telegr.*, 1928, **32**, 18-22, 65-70, 86-93.

After a brief treatment of the development of amplifiers for audio-frequencies, a detailed discussion is given of the electrical characteristics of audio-frequency amplifiers with resistance coupling between the various stages and the bearing of these characteristics on the acoustic qualities of such amplifiers. The second and third parts of the paper deal fully with the power transformer, the grid transformer, step-by-step volume control, automatic volume control, and practical applications of the results given.

3.2.3. TRANSFORMER AND CHOKE-CAPACITY-COUPLED AUDIO-FREQUENCY AMPLIFIERS

1. **Valve Amplifiers:** M. LATOUR: *Bull. Soc. franç. Élect.*, 1919, **9**, 465-84.

(See Abs. 3.1.2/1.)

2. **Effect of Transformer Resonance:** K. MÜHLBRETT: *Arch. Elektrotech.*, 1920, **8**, 365-90.

(See Abs. 3.3/3.)

3. **The Design of Inter-valve Transformers:** J. K. CATTERSON-SMITH: *Electrician*, 1920, **85**, 388-90, 414-6.

(See Abs. 3.3/1.)

4. **Notes on the Theory and Calculation of Audio-frequency Valve Magnifiers:** J. K. CATTERSON-SMITH: *Radio Rev.*, 1920, **1**, 473-80.

(See Abs. 3.3/2.)

5. **The Conditions for Distortionless Low-frequency Amplification:** F. M. COLEBROOK: *Experimental Wireless*, 1924, **1**, 448-53.

This is an early and very elementary paper calling attention to the fact that it is an essential condition of distortionless amplification that the anode-circuit impedance, if variable with frequency, must be large compared with the internal slope resistance of the valve over the whole required frequency range.

6. **Valve Characteristics:** E. GREEN: *Experimental Wireless*, 1926, **3**, 402-6, 469-76.

(See Abs. 3.2.1/6.)

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7. **Speech Amplifier Design :** N. W. McLachlan : *Wireless Wld.*, 1926, **18**, 44-48, 95-98, 137-42, 741-46.

This paper deals in a comprehensive manner with the problems of speech amplifier design. In the first section, the frequency-response characteristics of choke and transformer-coupled audio-frequency amplifiers are examined, and the requirements for uniform amplification are stated. It is pointed out that the predicted behaviour is only applicable to periodic waves, and that the many transient wave-forms delivered to the amplifier give rise to complex effects which cannot be accurately analysed. Since the response characteristic connecting radiated energy at the transmitter and acoustic input is not independent of frequency, it does not follow that the overall characteristic of the complete amplifier should show equal amplification at all frequencies. Circuit arrangements to produce a variable received frequency-amplification characteristic are given and their properties examined for particular cases. In the second part, details are given of a two-stage amplifier employing one resistance-capacity stage and one transformer stage in which the amplification-frequency characteristic can be varied to suit the constants of the loud speaker. In the last part, a simple analysis is given of the properties of the transformer coupled amplifier, and it is shown that a high impedance valve of about 40,000 ohms internal resistance with a transformer of a moderate inductance, say 20 henries, gives a diminished amplification at both high and low audio-frequencies, whereas a low impedance valve gives a good amplification at the low frequencies but there is a tendency for the higher frequencies to be accentuated.

(See also Abs. 3.3/12, 3.4/7.)

8. **Comparison between Transformer and Resistance Amplifiers :** M. von Ardenne : *Jahrb. drahtl. Telegr.*, 1926, **27**, 167-9.

An investigation of the amplifying properties of a transformer-coupled amplifier and a resistance-capacity-coupled amplifier, at a frequency of 1,000 p.p.s. To make a definite comparison possible, a formula is found for resistance-capacity-coupled amplifiers corresponding to the Barkhausen formula for amplifiers with transformer coupling.

9. **Low Frequency Inter-valve Transformers :** P. Willans : *J. Instn. Elect. Eng.*, 1926, **64**, 1065-88 ; and *Experimental Wireless*, 1926, **3**, 435-9.

(See Abs. 3.3/14, 4.4.2/7.)

10. **On the Values and the Effects of Stray Capacities in Resistance-Coupled Amplifiers :** M. von Ardenne and W. Stoff : *Proc. Inst. Radio Eng.*, 1927, **15**, 895-901.

(See Abs. 3.2.2/12.)

11. **A Plea for Choke-Coupling :** A. L. Sowerby : *Wireless Wld.*, 1927, **21**, 641-2.

The author examines the possible amplification obtainable by the use of large chokes in place of resistances, and shows that if a choke of 300 henries is used in conjunction with a valve having an internal resistance of 70,000 ohms, a satisfactory response characteristic can be obtained.

His analysis, however, neglects all shunting effects due to the input impedance and grid leak of the following stage, and it is unlikely that such results would be obtained in practice.

12. **Audio-frequency Transformers :** J. M. Thomson : *Proc. Inst. Radio Eng.*, 1927, **15**, 679-86.

(See Abs. 3.3/17.)

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13. **Theory of Terminal Amplification :** M. von Ardenne : *Jahrb. drahtl. Telegr.*, 1927, **30**, 116-23.

The author investigates the problem of the correct anode and grid voltages for distortionless reproduction with inductive loud speakers operated at maximum power output. The $e_g - i_a$ characteristic of an inductively loaded valve is an ellipse ; if this is enclosed by a rectangle the sides of which are parallel to the co-ordinate axes, the slope of the rectangle is also the dynamic slope of the valve for the given load. By means of this value, the correct grid and anode voltages can be obtained from equations which are given. The problems of loud speakers in series and in parallel and valves in parallel are considered, and a mathematical and a graphical solution of a particular example is given.

14. **Low Frequency Amplification with Choke Coupling :** H. Kafka : *Jahrb. drahtl. Telegr.*, 1928, **31**, 87-90.

The author investigates the frequency dependence of a " Hiler " coupling for audio-frequency amplification. This coupling consists of a choke in the anode circuit of the first valve, across which is a condenser in series with a second choke, to which the grid of the second valve is connected. The frequency dependence of the coupling can be influenced within wide limits by variation of the choke valves, so as to give a practically uniform amplification over a wide frequency range, or alternatively selective amplification of a definite frequency. It is claimed that this adaptability of the coupling is a great advantage over the transformer and resistance-capacity types of coupling.

15. **The Graphical Estimation of Low-frequency Choke Amplifier Performance :** W. A. Barclay : *Experimental Wireless*, 1928, **5**, 660-6.

Sowerby has utilised the anode voltage–anode current curve to derive the best values of resistance, &c., for given conditions of amplification by resistance-capacity-coupled amplifiers. A similar treatment of choke-coupled audio-frequency amplifiers is impossible owing to the increased number of variables, three instead of one. Indirect methods are, therefore, necessary, and the present paper describes a simple graphical construction, devised by the author to avoid the prohibitive labour of arithmetical work involved in the usual methods. The construction results in the finding of the centre and dimensions of the ellipse for greatest signal strength which will represent without distortion, at a given frequency, the working locus of a given choke, using a given H.T. voltage. The centre, when found, indicates the necessary grid bias to apply for the distortionless amplification of signals of the given strength and frequency. The paper includes a statement of the problem, a mathematical analysis and a proof of the construction.

3.2.4. Audio-Frequency Filter Circuits

(See Section 1.1.9.)

3.2.5.(*a*) The Power Stage and Power Amplifiers

1. **Vacuum Tube Amplifiers in Parallel :** R. V. L. Hartley ; *Proc. Inst. Radio Eng.*, 1921, **9**, 250-4.

This paper deals with the mathematical theory of operating a number of valves in parallel, i.e., all the grids connected and all the anodes connected together. Very often the valves are not uniform as regards their amplification constants and anode resistances. The effect of this is analysed, and the author concludes that the total power per unit input voltage that can be drawn from a group of valves is only the same in the common circuit as in

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individual circuits when the amplification constants of all the valves are equal. For this condition, it is not essential that the valve resistances should be equal. A large fraction of the total power is, however, available even if amplification constants and valve resistances are greatly dissimilar, an example being given.

2. **Design of Non-distorting Power Amplifiers :** E. W. KELLOGG : *J. Amer. Inst. Elect. Eng.*, 1925, **44**, 490-498.

The paper deals in an elementary manner with the problem of obtaining the maximum output from a given amplifier valve, while the distortion is reduced to a negligible amount. A valve can be rated for this purpose in terms of the watts output obtainable, when a sine wave voltage of as great an amplitude as can be advantageously utilised is applied to the grid. This maximum sine wave output is very much less than the rating of the same valve for oscillator purposes. Starting with a set of static characteristics for a given valve, the dynamic characteristic for any resistance load is readily plotted, and the power output and distortion can be read from the dynamic characteristic. The best load resistance is shown to be twice the internal anode resistance of the valve. If the supply voltage exceeds a certain value, the application of the rules just mentioned would lead to excessive heating of the anode, and therefore a different procedure is followed, calling for a greater grid bias and higher load resistance. There is an advantage in using low impedance valves. The balanced or push-pull circuit, while reducing distortion, will not make up for failure to operate the valves under proper conditions, nor will it greatly increase the permissible output per valve. The dynamic characteristic for a reactive load is not readily plotted, but for design purposes it is sufficient to determine the best operating conditions for a resistance load, and then make the impedance of the reactive load high enough to keep the anode current variations within the same limits as for the resistance load. An important application of the principles outlined here is the design of radio telephone transmitters where serious distortion results from overworking the modulator valves. For moderately deep modulation there should be from two to four modulating valves for each oscillator valve. Certain simple details of design are discussed in the closing paragraphs.

3. **The Output Characteristics of Amplifier Tubes :** J. C. WARNER and A. V. LOUGHREN : *Proc. Inst. Radio Eng.*, 1926, **14**, 735-57.

The effect of the valve constants on the amplification is first illustrated by three sets of curves showing the relation between the amplification constant of a valve and the actual amplification in terms of voltage, current and power. A simple method is given of determining the output of a valve when the amplitudes are large by making use of the family of anode voltage—anode current curves, and the condition that the load resistance should be twice the internal resistance for a maximum undistorted power output is proved. The calculation of the maximum undistorted output is next illustrated in detail, with and without regard to restrictions as to grid bias and anode current. The paper concludes with a discussion of the relation of the design of valves to the power output.

4. **The Performance of Valves in Parallel :** R. P. G. DENMAN : *Experimental Wireless*, 1927, **4**, 669-74.

A theoretical treatment is given of the general case of n batteries in parallel. From this, deductions are made for the case of valves in parallel. It is shown that, in general, better results would be obtained by having separate grid bias for each valve, but the practical difficulties of such an arrangement are obvious.

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5. **On Power Amplification without Distortion :** A. CLAVIER and I. PODLIASKY : *Onde Élect.*, 1927, **6**, 71-81.

Reference is made to previous work on this subject by E. W. KELLOGG (see Abs. 3.2.5/2), but it is treated in this article from a different aspect with a view to developing practical rules for use in the design of amplifiers and modulators. A formula is given which enables the optimum output resistance of a valve to be easily determined, given the anode dissipation permissible. The article concludes with some notes on the calculation of modulators and a classification of the more generally used commercial power-amplifying valves.

6. **The Last Stage :** A. P. CASTELLAIN : *Wireless Wld.*, 1927, **21**, 739-42.

This paper outlines the necessity for adequate power in the last stage and compares in a very elementary manner the relative advantages of connecting power valves in parallel as against their connection in " push-pull."

It points out that if the input grid swing is too large, paralleling valves does not improve the amplitude distortion.

7. **Theory of Terminal Amplification :** M. VON ARDENNE : *Jahrb. drahtl. Telegr.*, 1927, **30**, 116-23.

(See Abs. 3.2.3/13.)

8. **A New Method of Push-pull :** L. E. T. BRANCH : *Wireless Wld.*, 1928, **22**, 547-50.

(See Abs. 3.1.5/3.)

9. **Problems of the Final Stage of Amplification:** B. D. H. TELLEGEN : *Tijdschr. ned. Radiogenoot*, 1928, **3**, 141-60.

The action of the loud-speaker valve is considered as regards quality, amplification, volume and economy. If little frequency distortion is to occur, a valve whose internal resistance is high in comparison with the impedance of the loud speaker is required. This can be obtained, together with a large output, by putting a positive screen-grid between the control-grid and anode. To prevent secondary emission, a third grid between the screen-grid and anode is necessary which can be connected directly to the filament.

10. **Theory of Terminal Amplification :** M. VON ARDENNE : *Jahrb. drahtl. Telegr.*, 1928, **30**, 116-23.

Formulæ are developed for the least values of high tension and grid bias voltages required for distortionless reproduction, when inductive loud speakers are working at their maximum output with high emission valves. An examination is made of the conditions which exist when several loud speakers are connected in series or parallel, or when several valves in parallel are used to drive the loud speaker. Numerical and graphical solutions of the problem are given, together with curves giving the dependence of the useful output of the loud speaker on the internal resistance of the valve and the frequency.

(See Abs. 3.2.3/13.)

11. **The Theory of Push-pull :** N. W. MCLACHLAN : *Wireless Wld.*, 1928, **22**, 629-34.

In this paper the history of the push-pull circuit is briefly given and the conditions under which two valves will give undistorted output are examined. The necessity for the use of matched valves is pointed out and the two different methods of biasing the valves are examined. It is shown that the permissible undistorted output with the bias set so that the valves are almost at their rectifying points is double that obtainable by biasing to the mid-points of the linear portions of their characteristics.

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12. **On Banks of Paralleled Valves Feeding Resistive Loads without Distorting the Wave Form** : W. BAGGALLY : *Experimental Wireless*, 1928, 5, 315-21.

This paper gives numerical examples illustrating the manner in which the number of valves in parallel required to give a certain undistorted output can be calculated, and shows how to choose these valves to give a maximum output under varying conditions of high tension battery voltage.

13. **Distortionless Power Output of Electron Tubes** : A. FORSTMANN : *Jahrb. drahtl. Telegr.*, 1928, **31**, 45-50.

Formulæ are derived for the efficiency of a valve when used as a power amplifier and for the optimum resistance in the anode circuit. Relations are also obtained for determining the best operating conditions when both frequency and amplitude distortions are avoided.

14. **The Maximum Power Output of Amplifier Valves** : A. FORSTMANN and E. SCHRAMM : *Jahrb. drahtl. Telegr.*, 1928, **32**, 195-9.

Paying attention to the straight line conditions, the optimum values are determined for the external load-resistance, to produce maximum power output for a given anode voltage. The conditions as to the steepness of the characteristic and the amplification constant are also considered.

15. **Final or Power Stage of Amplifiers** : M. VON ARDENNE : *Experimental Wireless*, 1928, **5**, 556-64.

The author points out that when the input grid voltage to the last valve in an amplifier is assumed to be small, the maximum output is obtained when the differential resistance of the valve is made equal to that of the load at a selected frequency. If, on the other hand, the voltage which can be applied to the grid of the last valve without distortion is not limited, the maximum output is obtained when the resistance in the anode circuit is double the differential resistance of the valve. The subject is investigated graphically by means of the dynamic characteristics of the valve, and the result shows the derivation of the minimum anode voltage necessary for a given valve to deliver the maximum undistorted output to a given load. For purposes of speech amplification, the frequency is taken as 800 cycles per second, or as coinciding with the resonant frequency of the output load if this should be below 800 cycles per second.

16. **Voltage Limiter for Amplifiers** : O. KAPPELMAYER : *Jahrb. drahtl. Telegr.*, 1928, **32**, 25-26.

A simple limiting method employing a gaseous conducting valve for use with audio-frequency amplifiers is described, its function being to limit the alternating voltage of the grid of the last valve to a definite value, and so automatically prevent over-modulation of the valve.

3.2.5.(*b*). THE PENTODE

17. **The Pentode** : W. I. G. PAGE : *Wireless Wld.*, 1928, **23**, 7-9.

A description is given of the properties of the pentode, together with its characteristic curves. The essential properties are that of a tetrode, the valve possessing a high voltage factor, a high impedance and also a high mutual conductance. The kink in the normal characteristic of the tetrode, due to secondary emission from the anode to the screen, is eliminated by an earthed screen situated between the anode and the normal screen, thus making a valve with five electrodes.

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18. **Problems of the Final Stage of Amplification :** B. D. H. TELLEGEN : *Tijdschr. ned. Radiogenoot*, 1928, **3**, 141-60.

(See Abs. 3.2.5/9.)

19. **The Output Stage and the Pentode :** N. W. MCLACHLAN : *Wireless Wld.*, 1928, **23**, 30-33, 77-80, 113-6.

In this paper it is shown that the usual high and low note loss, which occurs when a coil-driven loud-speaker is operated from a low impedance power valve, may be considerably reduced by the use of the tetrode owing to its high internal resistance, and that it is necessary, if the full advantages of the valve are to be made use of, to use a loud-speaker impedance considerably higher than that required for a power triode. The pentode can take the place of an audio-frequency amplifying stage and a power valve, but care must be taken to reduce the input capacity due to grid leaks owing to the higher input impedance of this valve. A complete receiver embodying this new valve is described in detail.

(See also Abs. 3.2.6./6.)

20. **Permissible Grid Swing with a Pentode :** N. W. MCLACHLAN : *Wireless Wld.*, 1928, **23**, 584-6.

The author discusses the problem of grid swing as applied to a pentode, having in its anode circuit a reactance comparable in magnitude with the internal resistance of the valve. A sinusoidal voltage is applied to the control grid, the anode circuit containing an inductance and an a.c. feed meter, while an oscillograph is employed to delineate the wave-form distortion. The choke is shown to introduce serious wave-form distortion when the input voltage is well below the grid bias voltage.

21. **The Pentode and Power Amplification :** L. G. A. SIMS : *Wireless Wld.*, 1929, **24**, 60-64.

The author points out that serious damage to the valve may result from the high voltages produced by injudicious manipulation, e.g., if the loud-speaker is disconnected from the pentode while its filament is alight.

3.2.6. RETRO-ACTION

1. **Faithful Reproduction in Radio Telephony :** L. C. POCOCK : *J. Inst. Elect. Eng.*, 1924, **62**, 791-807.

(See Abs. 3.4/4.)

2. **Influence on the Amplification of a Common Impedance in the Plate Circuits of Amplifiers :** J. ANDERSON : *Proc. Inst. Radio Eng.*, 1927, **15**, 195-212.

The manner in which a common impedance in the anode circuits of two or more valves in an amplifier affects the output is examined, and it is shown that, in the absence of special feed devices, distortion or oscillation may occur. A theory is developed to explain this action and by its use several typical amplifier circuits are analysed. The paper concludes with some general experimental evidence, together with a discussion of methods avoiding this action.

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3. Battery Eliminators, or Appliances for the Operation of Radio Receiving Apparatus by Energy derived from Electric Supply Mains : P. R. COURSEY and H. ANDREWS : *J. Inst. Elect. Eng.*, 1927, **65**, 705-26.

In the discussion of this paper various comments are made regarding the stability of radio receivers in which the high tension supply is obtained from a battery eliminator.

The desirability of correctly designing the receiver to suit the eliminator is clearly demonstrated.

4. Back-coupling in Low Frequency Amplifiers : M. G. SCROGGIE : *Wireless Wld.*, 1927, **21**, 782-6.

An investigation is made into departures from the desired level amplification in audio-frequency amplifier units. In a multi-stage resistance-capacity-coupled amplifier the grids and anodes have opposite voltage phases, and consideration shows that the resistance common to all the anode circuits will weaken the amplified signal in an odd number of stages (inter-valve couplings) and strengthen it, and possibly cause oscillation, in an even number of stages. Measurements on amplifiers with a slide-back valve-voltmeter show the disturbing effects of battery resistance and battery eliminator impedance on the characteristic curves of amplifiers. The possibility of balancing out defects in other apparatus, such as the loud speaker, is discussed.

5. Battery Resistance and Distortion : W. I. G. PAGE : *Wireless Wld.*, 1928, **22**, 439-44.

(1) A choke filter output where one terminal of the loud speaker is connected to earth, or a choke filter-fed transformer, should be used with a receiver operating a loud speaker.

(2) Anode feed resistances should be used for every valve, but the last is the most essential if choke filter output is not used.

(3) Push-pull output, if the bridge balance is maintained, will prevent retro-action from the output valves.

(4) Output transformers do not tend to prevent audio-frequency oscillations.

(5) Choke filter output, where one loud speaker lead is taken to the positive terminal of the high-tension battery, does not prevent retro-action from the last valve.

(6) Reservoir condensers shunted across a high-tension battery give very little assistance in preventing retro-action.

(7) The useful life of a high-tension battery is considerably increased by the use of anode feed resistances ; 20,000 ohms is the lowest value recommended.

Such a resistance could not be used with a large power valve, owing to the large D.C. potential drop which would exist across it ; 2,000 ohms is a more usual value.

6. The Output Stage and the Pentode: N. W. MCLACHLAN : *Wireless Wld.*, 1928, **23**, 30-33, 77-80, 113-16.

The general properties of the pentode are first discussed, after which follows a discussion of the relative performance of a moving-coil and a reed-driven loud speaker when operated from a pentode and a triode. It is shown that owing to the high internal resistance of the pentode the upper frequencies will be quenched, and that, owing to the high factor, the input impedance is large, and this shunting effect will produce a high-note loss. In conclusion, the retro-action due to battery coupling is briefly discussed and the design

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of a three-valve receiver is given employing a screened-anode radio-frequency stage rectifier and a pentode output stage, the load of which is such as to minimise the high-note loss referred to above. The coils used throughout are of the toroid form.

(See also Abs. 3.2.5/19.)

7. **Effect of Anode-Grid Capacity in Detectors and L.F. Amplifiers:** W. B. MEDLAM: *Experimental Wireless*, 1928, **5**, 545-55.

In view of the importance of the anode-grid capacity in valve circuits, a mathematical investigation of the equivalent circuits is made in order to arrive at a quantitative measure of the effects observed. Most important are the retro-action and input load. In an anode-circuit rectifier, the retro-action effect is most prominent and leads, in practical cases, to considerable distortion in the apparent depth of modulation of the side-bands. In audio-frequency amplification, amplitude distortion may be introduced, and there is generally attenuation in the high audio-frequencies. To eliminate the disturbances caused by the capacitative retro-action, neutrodyned input and output circuits are recommended. For rectifiers, a neutrodyned input eliminates both the input load and the retro-action. In audio-frequency amplification, push-pull resistance-capacity coupling with neutrodyning is advocated.

3.2.7. AMPLIFIERS USING SPECIAL VALVES

1. **Resistance-coupled Amplifier:** H. KRÖNCKE: *Wireless Wld.*, 1925, **17**, 395-8.

(See Abs. 3.2.2/4.)

2. **The Screened Valve in Low-frequency Circuits:** N. W. MCLACHLAN: *Wireless Wld.*, 1927, **21**, 536-8, 577-9.

It is shown that no advantage can be gained by the use of the present type of screen-grid valve with iron-cored audio-frequency transformers; in fact, the reverse is the case, as the frequency-response characteristic becomes very peaked. The low self-capacity and high impedance, however, make the valve of service for the intermediate-frequency stages of a supersonic-heterodyne circuit, giving high magnification per stage with considerable selectivity.

With the usual resistance-capacity coupling, the screen-grid valve requires prohibitively large values of anode voltage supply, and even then gives less amplification than the ordinary high magnification type of triode.

3. **A New Method of using Resistance Amplification with Screen-grid Valves:** J. J. DOWLING: *Experimental Wireless*, 1928, **5**, 61-62.

From a consideration of the anode-volts anode-current curves for the above valves, the author shows how a very great amplification can be obtained with resistance coupling by working on that part of the valve characteristic which has a negative slope, and therefore behaves in a similar manner to the dynatron. He points out that the method is limited to amplitudes of the order of a few hundredths of a volt. Details are given of an experimental circuit used to check the theory in which an actual stable amplification of 150 was obtained.

The method is of interest, but has little practical value, owing to the critical conditions which need to be satisfied, together with the severe input limitations.

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3.2.8. Screening of Audio-Frequency Amplifiers

1. **On the Electromagnetic Screening of a Triode Oscillator :** R. L. Smith-Rose : *Proc. Phys. Soc. Lond.*, 1922, **34**, 127-38.

(See Abs. 1.2.9/1.)

2. **Shielding of Electric and Magnetic Fields :** J. H. Morecroft and A. Turner : *Proc. Inst. Radio Eng.*, 1925, **13**, 477-505.

(See also Section 1.2.9.)

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3.3. AUDIO-FREQUENCY TRANSFORMERS, ANALYSIS AND DESIGN

THE early types of iron-cored transformers gave amplification-frequency characteristics which were very peaky, owing to the natural resonances of the windings which possessed a small inductance and a large self-capacity. The inter-valve transformer may be considered as possessed of inductance, capacity, and resistance which are not very variable with frequency or voltage, so long as the latter does not exceed a few volts on the secondary. For the purpose of analysis, the transformer is considered as an inductance, a resistance, and a condenser all in parallel. DYE (see Abs. 3.3/5) has shown that such transformers behave towards the circuit to which the primary winding is connected in a manner which can be represented graphically by a circle diagram. The circle diagram enables the effects of mutual and secondary capacity and a secondary shunt resistance to be clearly visualised. The diagram enables the voltage across the primary winding to be obtained graphically.

The grid-filament circuit of the valve to which the secondary winding of the transformer is connected may produce profound effects upon the primary impedance. The input impedance of this following stage can be represented for analysis as that of a capacity and a resistance in series, and this relatively low impedance is thrown across the secondary as a load. The effect of a large input capacity is detrimental, and a secondary shunt resistance produces a flattening of the frequency-response characteristic. Since this is usually desirable, the addition of a moderate shunt resistance to the secondary winding is often adopted to produce greater stability and a better amplification characteristic. The amplification will, of course, be lowered throughout, but not seriously if the shunt resistance is not smaller than $\frac{[\text{the ratio of turns}]^2}{5}$ megohms. A turn or two wound over the secondary and short circuited will produce a similar improvement. The following points of design are deducible from the theoretical analysis :—

(i) The primary inductance must be large if the low audio-frequencies are to be maintained. At frequencies below 300 p.p.s. an amplification as great as 75 per cent. of the maximum cannot be obtained on the step-up transformer unless the primary inductance is in the neighbourhood of 20 henries.

(ii) The shunt secondary capacity must be small, which implies that the input impedance of the following stage must also be small.

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(iii) The mutual capacity between the windings should be small.

(iv) If no damping resistance is used to improve the frequency characteristic, the secondary inductance should be as large as possible, and the secondary leakage inductance and self-capacity should be as small as possible.

By dividing the secondary coil into several sections interleaved with the primary, both the leakage inductance and self-capacity of the secondary are reduced, but there will be a limit above which it is inadvisable to increase the secondary inductance, since the shunting effect of its own self-capacity nullifies the effect of the greater ratio. Again, since the primary inductance needs to be large, the possible step-up effect is necessarily limited and is usually between 1 : 1 and 1 : 4.

The design of the iron-core presents many difficulties. In the first place, the effective cross-section of the laminated iron-core is not accurately known, since the depth of penetration of the field into the iron is dependent upon the frequency. Secondly, a wave-form distortion is produced by hysteresis and since the magnetisation current increases as the frequency decreases, the distortion is more serious at the low audio-frequencies of the order of 50 p.p.s. Furthermore, the distortion produced at low audio-frequencies gives rise to harmonics which fall well within the audio-band, whereas at higher audio-frequencies, the harmonics fall out of the audio-band. Thirdly, the losses in the iron core are of a complicated nature; the behaviour of iron excited by a multiplicity of frequencies is not yet understood. It is known that the magnetic behaviour of iron at 40 p.p.s. is influenced by the action of the magnetisation due to a higher frequency, say 2,000 p.p.s., operating conjointly, especially at low flux densities. Fourthly, it is difficult to find the value of the magnetic permeability of the laminated iron core at different frequencies and under various conditions of polarising flux. The differential permeability is usually given as 200–400 for iron and up to 1,200 for nickel-iron alloys. The effect of the direct anode component of the primary current is to reduce the incremental permeability of the transformer core and, therefore, the primary inductance and induced secondary voltage. Decreased signal strength tends to reduce the incremental permeability and therefore the ratio of the voltages. If the anode direct current is large, silicon-steel may give nearly as good a performance as nickel-iron, but if the anode current is normal, nickel-iron cores are much superior at the lower frequencies.

The development of the transformer has of necessity been coupled with methods of measuring its performance, since certain factors which affect its behaviour are not readily calculable. There is a difference of opinion regarding the best method of expressing the behaviour of a transformer. One opinion is that the trnasformer

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is best measured as a unit by means of a bridge, and that if the constants of the transformer are thus known, the performance of a stage consisting of this transformer and any valve can be calculated. The second opinion is that bridge methods of measurement are unsatisfactory, since it is the amplification of the stage that is required and not the individual performance of the transformer. It is maintained that the complex effects due to input impedance and retro-action cannot be imitated by the use of a bridge measurement. In either case, however, the measured performance of the stage is only an indication of its properties, since the input impedance effects will modify the stage performance, and with two or three stages in cascade, the overall performance may differ very appreciably from the product of the stage performances, due to retro-action effects.

It is now appreciated that the transformer design must take into consideration the type of valve to be used with it. A high impedance valve gives a peaky characteristic for the stage, the best results being obtained with low impedance valves, though the superiority of the low impedance valve is less marked if used with high impedance transformers than if used with low impedance transformers. By suitable design, it is possible almost completely to damp out the primary resonance or "first resonance point" and to adjust the second resonance point to occur at about 4,000 p.p.s. By this means the upper audio-frequencies are maintained, and if the transformer stage is used in conjunction with a resistance-capacity coupled stage, this effect may be advantageous. The large primary inductance is now obtained by a larger iron section rather than by an increased number of primary turns. By this means the self-capacity of the secondary for a given turn ratio is far less than would be obtained with a larger number of turns and a smaller section of core.

BIBLIOGRAPHY

1. **The Design of Inter-valve Transformers**: J. K. Catterson-Smith: *Electrician*, 1920, **85**, 388-90, 414-6.

This paper deals with the design of iron-cored transformers for use with audio-frequency valve amplifiers for 500 to 1,000 cycles per second.

The ideal case is first considered, in which the transformer is assumed to have no magnetising current or losses. By resolving the primary resistance into a resistance in series with the secondary by well-known transformer theory, the following formula for the step-up ratio of one stage is obtained

$$K = \frac{\mu n R_g}{R_g + m^2 R_a}$$

where

K = the stage voltage factor,
μ = the valve voltage factor,
n = the step-up ratio of the transformer,
R_a = the internal resistance of the valve used with the transformer, viz. $\frac{dv_a}{di_a}$ and R_g = the secondary load resistance due to the following valve.

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The practical case is then considered where, unless the reactance of the transformer primary is large, a magnetising current flows and causes a change in the phase of the anode current. By ignoring the copper and iron losses, the stage factor is shown to be given by

$$K = \frac{\mu n}{\sqrt{\left[1 + \left(\frac{R_a}{Z}\right)^2 + \frac{2n^2 R_a}{R_g}\right]}},$$

where

$$Z = \left[\left(\frac{n^2}{R_g}\right)^2 + \left(\frac{1}{\omega L}\right)^2\right]^{-\frac{1}{2}}, \quad \omega L \text{ being the reactance of the primary.}$$

From this expression it is seen that the reactance of the primary must be large if the stage factor is to approach the value μn at low audio-frequencies.

The design of transformers is then described, and the necessity for using large reactance makes it advisable to adopt a closed iron magnetic circuit with a relatively high inductance per turn in order to minimise the capacity effect of the windings. The " shell " type with circular windings is recommended, and since the magnetising current is very small, a large number of turns is needed to give a relatively low permeability, expressions for which are derived. The use of such low densities permits iron losses to be neglected.

Transformers for two particular cases are then calculated, and one of these is illustrated. Finally, the use of auto-transformers for ratios of less than 3 or 4 is mentioned, and their application as telephone transformers indicated.

Although indicating the main features of design, the paper cannot be considered as representing actual conditions due to its neglect of many loading effects.

2. **Notes on the Theory and Calculation of Audio-frequency Valve Magnifiers :** J. K. CATTERSON-SMITH : *Radio Rev.*, 1920, **1**, 473-80.

This paper deals with difficulties that arise through the self-capacity of the windings of transformers used in valve amplifiers. The theory of the valve with transformer coupling is developed and formulæ are derived for use in calculating the windings and the size of the iron core. A first approximation for the capacity between layers is given. For a coil of mean radius r, of axial length b, with $(n + 1)$ layers, t cm. being the distance between layers and k the specific inductive capacity of the dielectric, it is found that the self-capacity C_s is given by $C_s = 1 \cdot 1\ krb/4nt\ \mu\mu F$.

It is not claimed that this formula is exact, but it gives a result usually in fair agreement with the value found experimentally.

3. **Effect of Transformer Resonance :** K. MÜHLBRETT : *Arch. Elektrotech.*, 1920, **8**, 365-90.

The subject is treated both theoretically and practically. The types of transformers are subdivided into :—

(1) Those lying between the receiving aerial and the first valve.

(2) Inter-valve transformers between anode and grid circuits of successive amplifying valves.

(3) The end transformer between the anode circuit of the power valve and the telephone.

The effect of transformer resonance is next considered for voltage and current resonance, the transformer being treated by means of an equivalent circuit, comprising an inductance and resistance in series, with a parallel circuit possessing inductance and capacity. Details are given regarding

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the practical methods of investigation of transformers carried out by the author on transformers of British, French, German and American manufacture.

A third section deals with investigations of transformers in conjunction with a valve, a large number of resonance curves being illustrated. The effect of a capacity in parallel with the transformer is dealt with, and also the effect of a short-circuit on the transformer windings. Finally, a summary of the test results is given with suggestions for improvements based on the decrease of damping, and the reduction of self-inductance and capacity of the windings, so far as practical considerations permit.

4. **Dimensions of the Choke Circuits in Radio-telegraphy** : J. BETHENOD : *Radioélectricité*, 1921, **1**, 629-32.

This paper deals with the two following problems :—

(1) In a given iron-cored choke coil to determine the number of turns for which the self-induction is a maximum, if the direct current flowing in the turns is given.

(2) In a given iron-cored choke wound with a given number of turns through which a direct current flows, to determine the length of the air-gap for which the self-induction is a maximum. In both cases the alternating current is supposed to be small in comparison with the direct current.

The problems are dealt with graphically by using the magnetisation curves and the limits of application of the method are indicated.

5. **The Performance and Properties of Telephonic-frequency Inter-valve Transformers** : D. W. DYE : *Experimental Wireless*, 1924, **1**, 691-9 ; **2**, 12-21, 74-84.

One test commonly applied to inter-valve transformers is to measure the effective voltage amplification given by a valve and transformer. This test, though very valuable as a measure of the overall performance of a transformer-coupled stage, does not give sufficient information, and further measurements are necessary if the transformer action is to be analysed. Methods are described for measuring the effective primary inductance and resistance of audio-frequency transformers. It is shown that such transformers behave towards the circuit in which the primary winding is connected as though consisting mainly of an inductance, a condenser, and a shunt resistance in parallel. Such a combination has effective reactances and resistances of which the corresponding values at various frequencies give a curve which approximates very closely to a circle. The circle diagram enables the effects of mutual and secondary capacity and a secondary shunt resistance to be clearly visualised. The diagram enables the voltage across the primary winding under given conditions to be determined graphically. It is then shown how the secondary inductance and self-capacity may be found. The effects of the foregoing quantities on the ratio of the transformer are examined both experimentally and theoretically, and it is shown that the ratio bears a close relationship to the coefficient of coupling. The grid-filament circuit of the valve to which the secondary winding of the transformer is connected may produce profound effects on the primary impedance, according to the conditions prevailing in the circuit. It is usually desirable to apply to this grid a steady negative E.M.F. with respect to the filament of such a value that there can never be a thermionic grid current. Under these circumstances, the grid-filament circuit becomes of very high impedance, representable as a small capacity and a resistance in series across the secondary winding, and does not seriously react on the primary impedance. Resonance conditions are fully dealt with. The methods given are applied to the

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determination of the amplification factor of the transformer, and the following conclusions are given :—

(i) Inter-valve transformers may be considered as possessed of inductance, capacity and resistance, which are not very variable with frequency or voltage, so long as the latter does not exceed a few volts on the secondary.

(ii) The addition of a moderate shunt resistance to the secondary winding will, in general, produce greater stability and a flatter amplification curve than will exist in its absence. The amplification will be lowered throughout, but not seriously if the shunt resistance is not smaller than the $\frac{[\text{ratio of turns}]^2}{5}$ megohms.

A turn or two wound over the secondary and short-circuited will produce a similar improvement.

(iii) At low audio-frequencies (below 300 cycles per second) an amplification as great as 75 per cent. of the maximum cannot be obtained on a step-up transformer, unless the primary inductance is greater than 20 henries.

(See also Abs. 4.4.2/2.)

6. **Telephone Transformers :** W. L. CASPER : *J. Amer. Inst. Elect. Eng.*, 1924, **43**, 197-209.

(See Abs. 4.4.2/3.)

7. **Inter-valve Transformers :** C. H. NAYLOR : *Elect. Rev. Lond.*, 1925, **96**, 589-91, 631-3.

A purely theoretical examination is undertaken into the properties of the inter-valve transformer and its accompanying valve ; the method adopted is that of graphical vectors. The conclusions arrived at are :—

(i) The amplification of the lower frequencies will be considerably improved by increasing the magnetic impedance of the primary, either by increasing the number of turns or the quantity of iron. Increasing the primary turns, however, entails increasing the secondary turns to maintain the ratio.

(ii) The amplification at high audio-frequencies is improved by reducing the capacity of the secondary. For a given capacity in the secondary, the amplification characteristic will be more constant with frequency variation for moderate ratios of transformation than with extreme ratios. Amplification at the higher frequencies will be greater if the transformer follows a valve having a low internal resistance, the only disadvantage being that low resistance valves have a large continuous current component, and the magnetic induction in the iron is increased accordingly ; the value of μ falls somewhat, causing lower impedance, and the amplification at the lower frequencies is slightly reduced. The low resistance valve, however, improves the high audio-frequency amplification much more than it adversely affects amplification on the lower frequencies.

8. **Resonance Capacities of Transformers :** R. HIECKE : *Elektrotech. u. Maschinenb.*, 1925, **43**, 97-98.

The effects of the capacities between windings, and between windings and core, self-capacity, etc., are discussed for audio-frequency amplifier transformers. A formula for resonance conditions is obtained, involving the above capacities, and particular cases are considered.

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9. **Distortion in Amplifiers :** G. W. SUTTON : *Wireless Wld.*, 1925, **17**, 156-9.

(See Abs. 3.1.2/3, 4.4.1/5.)

10. **The Impedance of Audio-frequency Amplifier Transformers :** S. CHIBA and S. KITTA : *J. Inst. Elect. Eng. Jap.*, 1925, No. 445, 733-8.

(See Abs. 4.4.2/5.)

11. **Audio-frequency Transformer Characteristics :** T. SPOONER : *Elect. J.*, 1926, **23**, 367-75.

The effect of the direct anode current of the associated valve is to reduce the incremental permeability of the transformer core and, therefore, the primary inductance and induced secondary voltage. Decreased signal strength tends to reduce the incremental permeability and therefore the ratio of the voltages. If the anode direct current is large, silicon steel may give nearly as good a performance as nickel-iron, but for ordinary voltage amplifiers nickel-iron is much superior at the lower frequencies. A simple method of test for the voltage amplification characteristics of the complete transformer is given, together with some results of such tests. As a voltage amplifier, when used with a transformer, a low impedance valve is superior to a high impedance valve ; the best performance is obtained if reduced anode and grid voltages are applied to the valve. The superiority of the low impedance valve is less marked if used with high impedance transformers than if used with low impedance transformers.

12. **Speech Amplifier Design :** N. W. MCLACHLAN : *Wireless Wld.*, 1926, **18**, 741-6.

The author discusses various problems in connection with the design and manufacture of high-inductance transformers. Distortion arising from hysteresis losses and saturation of the iron circuit is also considered. The author concludes with a brief statement of the advantages of the high-inductance transformer.

(See also Abs. 3.2.3/7, 3.4/7.)

13. **Transformers for Telephone Amplifiers :** R. HIECKE : *Elektrotech. u. Maschinenb.*, 1926, **44**, 49-57.

The author points out that, while some of the fundamentals of power transformer design can be applied to the telephone transformer, special conditions arise which make the design a compromise. The influence of the lower frequencies and the natural frequency of the transformer itself lead to distortion, while the internal resistance of the valve and the permeability of the iron require careful selection for optimum results.

The design of such transformers is very completely discussed.

14. **Low Frequency Inter-valve Transformers :** P. WILLANS : *J. Instn. Elect. Eng.*, 1926, **64**, 1065-88, and *Experimental Wireless*, 1926, **3**, 435-9.

An expression is developed for the voltage amplification of a valve and transformer and various particular cases are considered. The importance of low leakage inductance for the uniform amplification of the higher frequencies is emphasised. Measurements verify the theory and graphical methods are given for determining the transformer constants, and for determining the amplification ratio for frequencies beyond the range of measurement. A transformer design is described in which low leakage inductance and self-capacity are ensured by a sectionalised and spaced construction, and an amplification curve for this transformer is given.

(See also Abs. 4.4.2/7.)

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15. **Amplifier Transformers :** L. MÜLLER : *Arch. Elektrotech.*, 1926, **16**, 219-50.

A comprehensive theoretical discussion is first given. This is divided into five sections, dealing with :—

(1) Coil capacities, iron and copper losses, leakage, etc.
(2) Resonance frequencies, fundamental equations.
(3) Short-circuit conditions.
(4) Transformation ratio.
(5) Amplifier transformers with short-circuited windings.

A description is then given of experimental work on 22 transformers of three types, one type having a closed iron circuit, a second having a straight core of iron wires and enclosed in a screen consisting of soft iron with copper outside, the iron circuit being nearly closed, and the third having an open iron circuit. Experimental methods are fully described and many interesting results are illustrated graphically. Among these are diagrams showing the effects of internal short-circuits and of breaks in a transformer secondary winding.

(See also Abs. 4.4.2/8.)

16. **The Properties of the Circle Diagram for Telephonic Frequency Intervalve Transformers :** F. E. HACKETT : *Experimenal Wireless*, 1927, **4**, 601-4.

It is shown that to a very good degree of approximation the performance of an inter-valve transformer may be represented by a circle diagram. The paper also shows how the same results may be derived by the use of the crank or vector diagram.

This paper merely describes a simple method of analysing and expressing the properties of a circuit consisting of inductance, resistance and capacity in series. It is not applied to any particular properties of the transformer.

17. **Audio-frequency Transformers :** J. M. THOMSON : *Proc. Inst. Radio Eng.*, 1927, **15**, 679-86.

A method for calculating the amplification curve of an audio-frequency transformer is developed in terms of the usual constants, the distributed capacity of the coils, and the mutual capacity between primary and secondary represented by lumped capacities, while the transformer exciting current is neglected. The maximum amplification is obtained when the equivalent impedance of the transformer is very large compared with the anode impedance. There is no simple relation between the anode-circuit impedance and the amplification constant of the valve which will give the performance of the valve when used with audio-frequency transformers. The self and mutual capacities of the coils act as a by-pass at the higher audio-frequencies, the effect depending on the ratio of turns on the transformer.

18. **The Design of Transformers for Audio-frequency Amplifiers with pre-assigned Characteristics :** G. KOEHLER : *Proc. Inst. Radio Eng.*, 1928, **16**, 1742-70.

The author points out the essential characteristics required for the transformer-coupled stage. The equivalent A.C. circuit is solved, and the impedance and the voltage-amplification characteristics are explained. An examination of the theoretical results reveals difficulties which are encountered in design, and some methods of overcoming these difficulties are suggested. The effect of the continuous magnetic flux in the core of

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the transformer is illustrated, and a scheme is given for balancing out this flux.

The examination of the design details is very thorough, as is also the generalised analysis.

(See also Abs. 4.4.2/14.)

19. **Low-frequency Inter-valve Transformers :** R. JOUAUST : *Onde Élect.*, 1928, **7**, 437-45.

(See Abs. 4.4.2/15.)

20. **Characteristics of Output Transformers :** J. M. THOMSON : *Proc. Inst. Radio Eng.*, 1928, **16**, 1053-64.

The paper deals with the operating characteristics of the output transformer, which are developed in terms of the known loud speaker, valve and transformer constants. In the first part of the paper, the general formula for the loud speaker current is developed, and the effect of varying the transformer constants is shown. The turn ratio of the transformer for maximum loud speaker current is considered in relation to the commonly used impedance ratio formula. The limitation of the impedance ratio formula is then pointed out and limits set for its general use. The general form of the current-frequency characteristic for experimental horns and dynamic cone loud speakers is then obtained, and a general method for matching the loud speaker to the output valve is given. In the latter part of the paper, curves are given to show the results obtained in the mathematical part of the paper. The curves also include the results of tests made to check the fundamental formula. The effect of the turn ratio on the form of the current-frequency characteristic is shown, and a method of using the turn ratio of the output transformer to match the loud speaker to the output valve is given. A perfect transformer is compared with a good commercial transformer, and the general effect of the leakage inductance and the self-capacity of the transformer is shown.

21. **Transformer Coupling in Audio-frequency Amplifiers :** H. REPPISCH : *Jahrb. drahtl. Telegr.*, 1928, **32**, 22-24.

Formulæ are derived for the voltage relations of transformers used as coupling units between the various stages of audio-frequency amplifiers.

22. **Electrical Characteristics of Amplifiers for Broadcasting Transmissions and their Relation to Acoustics Quality :** H. RUKOP : *Jahrb. drahtl. Telegr.*, 1928, **32**, 18-22, 65-70, 86-93.

(See Abs. 3.2.2/27.)

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3.4. ANALYSIS OF DISTORTION IN AUDIO-FREQUENCY AMPLIFIERS

The four types of distortion which can occur in audio-frequency amplifiers can be classified as follows :—

(i) Frequency distortion.
(ii) Amplitude distortion.
(iii) Wave-form distortion.
(iv) Phase distortion.

Frequency distortion is obtained when all frequencies are not uniformly amplified. Since this type of distortion is so essentially bound up with the design of the amplifier, most of the references to this subject will be found in the Bibliography, sections 3.2.2, 3.2.3. The literature on the subject is profuse, and only in those cases where the paper applies entirely to frequency distortion is it inserted in this section. In resistance-capacity-coupled amplifiers, the frequency distortion takes the form of a cut-off below a certain low audio frequency and above a certain high audio frequency. A low audio frequency cut-off is obtained by too small a coupling condenser between the stages, whereas a high-note cut-off is produced by :—

(*a*) Too large an input impedance load due to the following valve.
(*b*) Too large a by-pass condenser across the anode resistance of the rectifier valve.
(*c*) Anode resistances having too large a self-capacity.

In transformer-coupled amplifiers, both the high and the low cut-off points are determined by the characteristics of the transformer and the valve. The amplification-frequency characteristic of a choke-capacity-coupled stage, in which the inductance is not less than 50 henries and the valve voltage factor is low, is better than for a transformer, but the step-up effect is, of course, absent.

The previous remarks refer to the frequency distortion which can occur in the amplifier. Frequency distortion may, however, be produced by the loud speaker output. Although little is definitely known about such distortion, owing to a lack of experimental data, the following general inferences may be made :—

(i) The natural frequency of the moving part of the output mechanism must be high and the mass small for good reproduction under steady-state conditions.

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(ii) The response of a system to a suddenly applied steady force may be regarded as a good indication of its behaviour towards the transients that will occur in speech and music.

(iii) For analytical purposes, the response-frequency characteristic is related to the response for transients as well as for the steady state.

(iv) The response of a system to a transient applied force depends upon the admittance expressed in a complex form, i.e., as a modulus with a phase angle.

The relation of response to input, i.e., frequency-distortion characteristic, is determined as a modulus of the mutual admittance, no account being taken of the phase angle. Experiments show that alterations of wave form due to alterations of phase do not materially affect the sound heard, and justify this statement for the steady state. Under transient conditions, however, the shape of the response characteristic may be greatly altered by phase distortion.

Various complicated systems of chokes and filter circuits have been developed to correct the poor frequency characteristic of the loud speaker. In the absence of any accurate information regarding the response characteristic of the loud speaker such " tone-correction " is at present purely empirical.

Amplitude distortion is obtained when the voltage output is not directly proportional to the input, or in other words, when the amplification value is not independent of the amplitude of the applied signal. Such distortion may be produced by operating the valve at a non-linear portion of its characteristic, either by using incorrect grid-bias or by applying too large a signal E.M.F., or by the use of poor inter-valve transformers in which saturation occurs.

In dealing with the amplitude distortion produced by an amplifier or a loud speaker mechanism, the physiological aspect of the matter must not be overlooked. The human ear is comparatively insensitive to small changes of pressure about a mean " loudness level," and experiments to detect what percentage change of intensity is just noticeable have been made. The percentage variation in sound amplitude required to produce a change in strength just noticeable by the human ear is given by some writers as 5 per cent. and by others as 30 per cent., but these figures are open to much question owing to the very divergent opinions of different individuals. If quantitative information is required regarding amplitude distortion perceptible to the human ear, some such figure representing the physiological limit of sensitivity is required. If there is no amplitude distortion, the ratio of output to input defines the frequency distortion, viz.—

$V_2 = G(f)V_1$ where $G(f)$ represents the frequency distortion and is a function of frequency alone.

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If amplitude distortion is present V_2 varies non-linearly with V_1 at a given frequency, viz., $V_2 = G(f)\ H\ (V_1)\ V_1$

Wave-form distortion occurs when the amplification process gives rise to frequencies which are not necessarily present in the original input wave-form. Such distortion can be produced by any or all of the following causes ;—

(i) Operating the valves beyond their normal linear dynamic characteristics.

(ii) Permitting the grid swing to make the grid positive during part of the periodic cycle, thus introducing grid current.

(iii) Operating the iron-core transformer on a non-linear portion of its B-H. curve. This can occur if the signal is very weak, in which case lower bend distortion is produced; or if the signal is too great, in which case distortion due to the upper bend is produced. The same effect occurs if the polarising current is such as to place the operating point on a curved portion of the B-H. curve.

(iv) Due to hysteresis and eddy-current effects in the iron-core, which give a non-linear relationship between the instantaneous input and output.

(v) Due to incorrect grid-biasing of push-pull output stages, or the use of unmatched valves, in which case the positive and negative half-cycles are dissimilar.

(vi) Due to grid charging effects which give rise to transient distortions when excessively large peak voltages occur.

With iron-core transformers, the distortion is more serious at low audio-frequencies than at high audio-frequencies, and the effect of negative grid bias is to reduce the magnitude of the lower harmonics and to increase the amplitude of the higher harmonics (see Abs. 3.4/9). There is an optimum value of the polarising current in an iron-cored transformer where the wave-form distortion is a minimum, but, even at this value, the distortion is not negligible. There is less possibility of wave-form distortion in resistance-capacity coupled amplifiers than in amplifiers employing a coupling arrangement involving the use of iron.

It is difficult to examine the distorting effects of an amplifier upon telephony signals owing to the transient nature of the potentials which cannot be expressed as periodic functions of time, but require an analysis involving the Fourier series or Heaviside's Expansion theorem (see Abs. 1.2.1/23, 15).

Fortunately, phase distortion in audio-frequency amplifiers does not affect the reproduced output, since the relative phases of the harmonics can be shifted without audible detection so long as the amplitudes and frequencies remain correct. For ordinary telegraphic and telephonic purposes, phase distortion is of no

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importance, but in special problems where two identical amplifiers are required, e.g., to operate the two pairs of plates of a cathode-ray oscillograph, the phase distortion may not be the same in both. Facilities for phase adjustment may then become necessary.

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1. **The Unilateral Dynamic Characteristics of Three-electrode Vacuum Tubes :** J. G. FRAYNE : *Phys. Rev.*, 1922, **19**, 629-51.

For the case of pure resistance anode circuit loads, the Van der Bijl parabolic relation between anode current and grid-voltage is expressed as a power series of $e \sin \omega t$, the impressed simple harmonic grid voltage.

The coefficients of the various harmonics involve R, the external resistance, the nth harmonic reaching a maximum value when

$$R = \frac{n-2}{3} \times R_0$$

where R_0 is the valve internal resistance. For the fundamental frequency the ***maximum energy output*** is secured when $R = 0{\cdot}81\ R_0$. The dynamic characteristic obtained by compounding the harmonics into a single curve approximates to a straight line as the resistance is increased. For the case of a pure inductance, the anode current is expressed as a Fourier series, the dynamic characteristic being a closed loop, the area of which is proportional to the energy in the inductance. The use of a condenser load gives a similar elliptical dynamic characteristic, but the phase angle of the various harmonics is shifted.

The effects on the anode current of varying the input grid voltage e, the static grid voltage E_c and the static anode voltage E_b, for a given value of resistance R or inductance L, and the effect of varying R or L with constant E_b, E_c and e, are determined and are shown in curves, together with the corresponding theoretical values. The results show that the equations predict the harmonic constituents of the anode current as high as the fourth for values of e up to 20 volts (the range for which the fundamental equation holds). For this range, the coefficients of the various harmonics in the equation are proportional simply to e^n and the fundamental becomes greater while the other harmonics diminish as we approach the straight portion of the static characteristic and as we increase the anode potential.

The frequency used was 200 kilo-cycles per sec., and consequently the valve capacities produced a negligible effect. The harmonics were picked out and measured by independent tuned circuits in the anode circuit, the constants of these circuits being accurately known. The agreement is very good up to 15 volts input with 200 volts on the anode.

2. **Distortion in Low-frequency Amplifiers :** S. O. PEARSON : *Wireless Wld.*, 1923, **13**, 3-6, 40-44.

The author gives a very elementary and popular discussion of various sources of distortion in amplifiers, particular attention being given to permeability and hysteresis effects in iron-cored transformers.

3. **Distortion in Radio-telephony :** H. A. THOMAS : *Wireless Wld.*, 1923, **13**, 225-9, 257-60, 291-3, 328-9, 356-7.

A theoretical examination of the different factors which can produce frequency and amplitude distortion is given, commencing with the microphone and passing successively through the modulating valves, oscillator, transmitting and receiving aerials, rectifier, amplifier and loud speaker.

Particular attention is given to the various types of microphone, and suggestions are given for the improvement in the design of this component.

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The effects of a lightly damped receiving aerial system or the use of excessive retro-action are examined, and it is shown that a severe cut-off of high audio-frequencies can be obtained. The various types of amplitude and frequency distortion which can occur in the valves, transformers and loud speaker are considered in some detail, and the wave-form distortion which can be produced by grid currents is mentioned. A simple vector diagram theory of the telephone receiver is given and the paper concludes with a discussion of comparative tests made between transformer coupled and resistance-capacity-coupled audio-frequency amplifiers.

4. **Faithful Reproduction in Radio-telephony :** L. C. Pocock : *J. Instn. Elect. Eng.*, 1924, **62**, 791-807.

An important paper upon the subject of distortion of speech in radio-telephony. Terms are defined such as articulation, perfection of reproduction, etc., to permit further study. Various types of distortion such as frequency distortion, amplitude distortion, etc., are analysed. Transient phenomena are considered, but only in relation to the simplest cases of the ironless circuit. A second part deals in a very general manner with the applications of these principles to amplifiers, retro-action, reflex circuits, receivers, loud speakers and microphones.

5. **The Conditions for Distortionless Low-frequency Amplification :** F.M. Colebrook : *Experimental Wireless*, 1924, **1**, 448-53.

(See Abs. 3.2.3/5.)

6. **Distortion in Radio-telephony :** E. K. Sandeman and N. Kipping : *Experimental Wireless*, 1925, **2**, 756-64, 811-19.

(See Abs. 4.5/6.)

7. **Speech Amplifier Design :** N. W. McLachlan : *Wireless Wld.*, 1926, **18**, 44-48, 95-98, 137-42, 741-46.

These three articles deal in a comprehensive manner with the problems of amplifier design. In the first article, frequency distortion due to audio-frequency chokes and transformers is considered ; in the second, details are given of a variable characteristic two-stage amplifier. The final article deals with the derivation of transformer characteristics. Various correction circuits are described which may be used to reduce distortion due to loud speaker resonances.

(See also Abs. 3.2.3/7, 3.3/12.)

8. **Transformers for Telephone Amplifiers :** R. Hiecke : *Elektrotech. u. Maschinenb.*, 1926, **44**, 49-57.

(See Abs. 3.3./13.)

9. **The Performance of Amplifiers :** H. A. Thomas : *J. Instn. Elect. Eng.*, 1926, **64**, 253-78.

The wave-form distortion occurring in audio-frequency amplifiers is examined experimentally by means of an Einthoven Galvanometer for frequencies up to 300 p.p.s., and by means of a Cathode Ray Oscillograph for frequencies above 300 p.p.s. The output wave form obtained is analysed harmonically and it is shown that :—

(i) The effect of increasing the output is to seriously modify the wave form. In general, the harmonics increase in magnitude far more rapidly than the fundamental.

(ii) The distortion is more serious at low audio-frequencies than at high audio-frequencies, and the harmonics may greatly exceed the fundamental.

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(iii) The negative grid bias reduces the magnitude of the second and third harmonics and introduces small amplitude harmonics of a higher order.
(iv) The form factor of the output wave remains sensibly constant.
(v) There appears to be a definite wave form distortion introduced by the transformer, and this is a minimum at a certain definite value of the polarising D.C. current.

The difference between an R.M.S. measurement of output and a measurement of the fundamental only taken by a resonant instrument may be as great as 20 per cent., and it appears that peak values give a truer indication of acoustic output than R.M.S. values.

(See also Abs. 1.2.7/6, 1.4.1/16, 4.2/8, 4.4.1/7, 4.5/7, 4.6/1.)

10. **Tone Correction in Low-frequency Amplifiers :** H. LLOYD : *Wireless Wld.*, 1926, **18**, 11-15.

The paper describes attempts to correct the effects of the non-uniform loud speaker response characteristic. This is done by the use of a complicated system of chokes, filter circuits and band-filters connected in the output circuit of the audio-frequency stages.

In view of the absence of exact acoustic response characteristics, the form that such a filter must take will depend upon the desired output characteristic ; divergent opinions on this matter will be obtained from different individuals due to temperamental and physiological reasons, and it appears doubtful whether such filtering devices can give better reproduction in general as distinct from changes which are considered to be improvements by particular individuals.

11. **Valve Distortion in Amplifiers :** M. VON ARDENNE : *Z. tech. Phys.*, 1927, **8**, 235-9.

The amplitude distortion arising from the curvature of the valve characteristics is investigated, together with the dependence of this curvature on the value of the anode impedance. The permissible amplitude distortion which is physiologically undetectable is about 5 per cent. for orchestral music.

It is doubtful whether such a figure has much significance owing to the difficulty of obtaining it in view of the serious divergence of opinion between individuals of different musical temperaments.

12. **Influence on the Amplification of a Common Impedance in the Plate Circuits of Amplifiers :** J. ANDERSON : *Proc. Inst. Radio Eng.*, 1927, **15**, 195-212.

(See Abs. 3.2.6/2.)

13. **Modulation in Vacuum Tubes used as Amplifiers :** E. PETERSON and H. P. EVANS : *Bell System Techn. J.*, 1927, **6**, 442-60.

These authors show that experimental results do not agree with the theories of CARSON and VAN DER BIJL (see Abs. 1.2.2/8, 2.3.2/6, 1.2.2/5). The cause of this discrepancy was found to lie in the neglect of the variation of the amplification factor with both anode and grid potentials. A new analysis was developed which allows of accurate calculation of the distortion to be expected.

14. **Sources of Distortion in Resistance Amplifiers :** M. VON ARDENNE : *Wireless Wld.*, 1927, **20**, 395-9.

The paper deals with the frequency distortion that occurs in multi-stage resistance-capacity-coupled amplifiers. The author considers that the

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frequency-response characteristic of the loud speaker is so bad that refinements in the characteristic of the amplifier are not needed. If the higher audio frequencies are only amplified one tenth as much as the lower ones, he states that even this reduction is not serious compared to the much more rapid fall in response of the loud speaker at high audio frequencies. The reduction in inter-stage capacities obtained by the use of " multiple-valve " resistance-capacity-coupled amplifiers of the Loewe type is discussed and the merits of these valves are examined.

The above statements with regard to the harmlessness of high note loss in resistance amplifiers appear to be an attempt to vindicate this type of amplifier at the expense of the loud speaker. The progressive improvement in the design of the loud speaker will probably invalidate this contention, if it has not done so already.

15. **Amplification and High Quality :** N. W. McLachlan : *Wireless Wld.*, 1928, **22**, 463-6.

In this article, the various sources of high-note loss in a broadcast receiving set are enumerated and the remedies suggested. They are, briefly, too sharp tuning of the aerial circuit ; too large a by-pass condenser for the anode resistance of the detector valve ; the effective capacity across, say, the audio-frequency transformer due to the input impedance of the power valve fed from it ; finally, the input impedance of the power valve as affected by the load in its output circuit. The distortion due to the detector valve is also mentioned.

16. **Microphone Amplifiers and Transformers :** H. L. Kirke : *Experimental Wireless*, 1928, **5**, 361-70, 443-51.

The author considers the various causes of distortion and how they may be avoided. Distortion of the frequency characteristic may be due to bad transformer design, the coupling condensers being too small in resistance-capacity-coupled amplifiers, too low a value of inductance in the anode choke, retro-action effects, and badly designed volume controlling apparatus. Distortion of the amplitude-magnification characteristic may be due to grid current in the valves, non-linear anode current characteristics, or to iron distortion in the transformers and chokes. Each of these is considered in some detail.

The paper is of a general nature only and does not examine any properties very fully.

17. **Permissible Grid Swing with a Pentode :** N. W. McLachlan : *Wireless Wld.*, 1928, **23**, 584-6.

(See Abs. 3.2.5/20.)

18. **Effect of Anode-grid Capacity in Detectors and L.F. Amplifiers :** W. B. Medlam : *Experimental Wireless*, 1928, **5**, 545-55.

(See Abs. 3.2.6/7.)

19. **On Banks of Paralleled Valves Feeding Resistive Loads without Distorting the Wave-Form :** W. Baggally : *Experimental Wireless*, 1928, **5**, 315-21.

(See Abs. 3.2.5/12.)

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SECTION IV—MEASUREMENTS

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4.1. CONSTANTS OF VALVES RELEVANT TO THE DESIGN OF AMPLIFIERS

THIS section deals with methods of measuring those constants of the valve which are of importance in the design of radio receivers. Citation is made only of those constants which affect the circuit behaviour as distinct from constants of the valve itself, such as the internal dimensions and physical conditions of the valve; these latter properties fall under the heading of Thermionics.

4.1.1. VALVE CHARACTERISTICS

The two most important constants which fundamentally affect the circuit properties are the internal resistance (sometimes spoken of as the "differential resistance") and the voltage factor. The former constant may be obtained from the static characteristic plotted between the anode voltage and the anode current, while the latter constant may similarly be obtained from the static characteristics as the ratio of the anode volts-anode current slope at the operating point to the anode current-grid voltage slope at the same operating point. Such a determination is accurate but not rapid, and an alternative method commonly employed is that due to MILLER (see Abs. 4.1.1/1). By means of a potentiometer, an alternating potential is applied to the grid, while, at the same time, a larger potential is supplied to the anode in antiphase; by varying the ratio of these potentials a telephone balance may be obtained, at which point the alternating component of the anode current is reduced to zero. The ratio of the potentiometer supplying these two potentials gives the voltage factor of the valve.

To obtain a rapid indication of the characteristic relating the anode potential to the grid potential, the Cathode Ray Oscillograph see Abs. 4.1.1/5, 3) has been used by SCHNEIDER and others. In this method, the grid of the valve is impressed with a pure sinusoidal wave from an oscillator at an audio-frequency, and this same E.M.F. is applied to one pair of oscillograph plates; the other pair are connected across a resistance in the anode circuit of the

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valve and give rise to a deflection of the beam proportional to the anode current. In this manner a curve was obtained giving the dynamic characteristic under any conditions. The advantage of this method is that the dynamic characteristic can be quickly obtained for any type of anode circuit load, the calculation of the dynamic characteristic from the static characteristics being tedious. In the case of a phase displacement, the form of the ellipse gives the phase angle as well as the stage factor.

4.1.2. "Schrot" Effect

Considerable experimental work has been devoted to the determination of the "Schrot" voltage produced across a tuned circuit in the anode circuit of a single stage, when no input is applied. Owing to the fundamental physical nature of the "Schrot" effect, the method has been used to determine the electro-magnetic value of the electronic charge, it being claimed that, if sufficient care is taken, the method rivals the "oil-drop" method. Hartmann (see Abs. 4.1.2/1) gives two methods: one consists of a bridge method of measuring the "Schrot" voltage, the bridge being excited by a source having the same frequency as the natural frequency of the tuned circuit and a balance obtained; the other consists of a substitution method, the "Schrot" voltage being applied to an amplifier which is calibrated by an oscillator feeding a potentiometer of known resistance and carrying a known current. Hull and Williams (see Abs. 4.1.2/4) have improved this latter method by improving the screening which is so essential when such small voltages are being measured. Discrepancies between the observed "Schrot" effect and predictions from theory have been found by several experimenters; one noteworthy effect is that there appears to be a slight diminution in the ratio of the observed effect to the calculated effect as the anode current is increased. The "Schrot" effect is considered more fully in Section 1.2.6.

4.1.3. Inter-Electrode Capacities and Input Impedance

Modern progress in the design of amplifiers is due largely to devices which compensate for the effects due to the inter-electrode capacities of the valve and the measurement of the values of these capacities constitutes a major necessity. The bridge method is eminently suitable for this purpose and Hartshorn has published several papers in which the application of the Schering Bridge to this problem has been adopted (see Abs. 4.1.3/3, 4, 6, 11). The measurement is made by obtaining the capacity between different electrodes, by which means the three independent capacities are measured in various combinations, a deduction therefrom being necessary to obtain the independent values. Such a method is capable of measuring capacities of the order of 1 micro-micro-farad.

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In bridge measurements of this kind, errors can be produced due to the capacities between the bridge arms to earth acting as shunts to the small capacities being measured; a Wagner earth is used to eliminate this effect. WALSH (see Abs. 4.1.3/10) has developed a bridge method which measures the capacity between any two electrodes directly. A standard form of capacity bridge is used in which the grid and anode are connected to the two terminals provided for the unknown capacity. The filament is connected to the common point of the two ratio arms, which is earthed. By this means, the filament-anode capacity is effectively inserted across the detector and does not affect the balance. The filament-grid capacity is in parallel with one of the ratio arms, and its effect can be balanced by a resistance inserted in series with the standard capacity.

Since a considerable part of the valve capacity is through the vacuum and does not change with frequency, the change of capacity with frequency due to the dielectric cannot be a large proportion of the total, and therefore audio-frequency capacity measurements can usually be applied to computations at radio-frequencies without serious error. However, when it is desired to take the conductance into account, radio-frequency measurements are preferable. HARTSHORN has shown that a large portion of the total capacity is due to the "pinch" of the valve, and also that the capacities are never pure, i.e., the capacity is associated with a parallel conductance, which conductance depends upon the frequency. Considerable variations from the nominal phase angle of 90° have been noticed, such variations being dependent upon the load in the anode circuit, the anode and filament voltages, and the frequency.

Owing to the difficulties associated with bridge measurements at radio frequencies, substitution methods have more commonly been employed to measure small capacities at such frequencies (see Abs. 4.1.3/7, 2). The current passing through a small capacity can be indicated by some amplifying device, and a known variable condenser can be substituted and varied until the same current is obtained. Such methods usually have a limited accuracy due to the difficulty of lead capacities.

BIBLIOGRAPHY

4.1.1. VALVE CHARACTERISTICS

1. **A Dynamic Method for Determining the Characteristics of Three-electrode Vacuum Tubes :** J. M. MILLER : *Proc. Inst. Radio Eng.*, 1918, **6**, 141-8.

A description is given of a method for determining the amplification constant and internal resistance of a valve by the potentiometer method. An alternating potential is fed to the grid and at the same time another potential is fed to the anode in anti-phase by regulating a potentiometer to balance the current changes due to each potential. Silence in the telephones

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is obtained, and the ratio of the two potentials supplied gives the voltage amplification factor of the triode.

2. **The Dependence of the Amplification Constant and Internal Plate-circuit Resistance of a Three-electrode Vacuum Tube upon the Structural Dimeusions :** J. M. MILLER : *Proc. Inst. Radio Eng.*, 1920, **8**, 64-73.

In a previous paper (see Abs. 4.1.1/1) the author pointed out the importance of the above two factors in determining the behaviour of valves as amplifiers, and outlined a simple and direct method of measuring these quantities. In the present paper, formulæ based on theoretical considerations are derived showing the dependence of these quantities on the dimensions and spacing of the three electrodes. A set of curves based on the derived formulæ allow of the rapid pre-determination of the amplification constant when the physical dimensions of the valve are known. The curves are shown to give results agreeing with experiment for a number of different types of valves, and the internal anode-circuit resistance formula is shown to be valid under certain conditions and deviations in specific instances are explained.

3. **Oscillographic Study of Electron Tube Characteristics :** E. L. CHAFFEE : *Proc. Inst. Radio Eng.*, 1922, **10**, 440-50.

A special oscillograph is used to record simultaneously the anode current and the grid potential. By superimposing a low audio-frequency variation of voltage on the regularly increasing grid voltage during the photographing of the oscillogram, corresponding audio-frequency variations are produced in the anode current, the amplitudes of which are proportional to the amplification of the valve at the particular setting.

Rectifier action can similarly be studied by applying a rising radio-frequency voltage to the grid, and periodically interrupting it at a low audio-frequency.

The study of valves containing gas, and the recording of kinks in their characteristics, is particularly facilitated by the use of the apparatus described.

4. **On the Theory of the Triode Valve Amplifier and the Measurement of the Characteristic Constants of the Triode Valve :** H. NUKIYAMA and S. KITTA : *J. Inst. Elect. Eng. Jap.*, 1925, No. 445, 664-78.

By measuring the amplification factor by an alternating current potentiometer, these three valve constants can be determined. A direct potentiometric method of determining the amplification factor is also given.

(See also Abs. 1.2.1/18.)

5. **Use of an Oscillograph for Recording Vacuum-tube Characteristics:** W. A. SCHNEIDER : *Proc. Inst. Radio Eng.*, 1928, **16**, 674-80.

This paper describes the methods involved and the results obtained in using a Cathode Ray Oscillograph for plotting valve characteristics photographically. The grid of the valve is supplied with a pure sinusoidal E.M.F. wave from an oscillator at a frequency of 120 p.p.s., and this same E.M.F. is applied to one pair of the oscillograph plates. The other pair are connected across a resistance in the anode circuit of the valve and give rise to a deflection of the beam proportional to the current. In this manner, a curve is obtained with rectilinear axes which can be photographed. The negative resistance action of triodes when large alternating E.M.F's. are applied to the grid is clearly shown. Both static and dynamic characteristics can be readily recorded by this method.

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4.1.2. "SCHROT" EFFECT

1. **Determination of the Elementary Electrical Quantum from the "Schrot" Effect :** C. A. HARTMANN : *Ann. Physik.*, 1921, **65**, 51-78.

In this paper a rough measurement of the value of the "Schrot" voltage, as predicted by SCHOTTKY, is given. Two methods of measurement are described. In the first case, the weak "Schrot" voltage existing across the tuned anode circuit of a diode is balanced by means of a bridge excited by a source having the same frequency as the natural frequency of the tuned circuit; in the second case, the "Schrot" voltage is applied to an amplifier and a certain output is obtained, after which a similar output is given by a local signal applied to the amplifier by an oscillator feeding a potentiometer of known resistance and carrying a known current. This current is measured by a thermo-junction. Elaborate screening is resorted to.

(See also Abs. 1.2.6/2.)

2. **The Schottky Effect in Low-frequency Circuits :** J. B. JOHNSON : *Phys. Rev.*, 1925, **26**, 71-85.

(See Abs. 1.2.6/5.)

3. **The Theory of the Schroteffekt :** T. C. FRY : *J. Franklin Inst.*, 1925, **199**, 203-20.

(See Abs. 1.2.6/6.)

4. **Determination of Elementary Charge E from Measurements of Shot-effect :** A. W. HULL and N. H. WILLIAMS : *Phys. Rev.*, 1925, **25**, 147-173.

SCHOTTKY first pointed out that if the electrons evaporate independently of each other, probability fluctuations of current are to be expected. These fluctuations were observed and roughly measured by HARTMANN. In this paper, measurements are made using a much higher frequency (750×10^3) and a reliable arrangement for measuring the small R.M.S. voltages (of the order 10^{-4}) due to the effect. The method involves the measurement of the alternating current excited in a tuned circuit by the variations in the electron current through a valve, the Schottky equation for the mean square current being

$$I^2 = i_0 e/2 \text{ R.C.}$$

where i_0 is the thermionic current,

R and C the resistance and capacity of the tuned circuit, and
e the electronic charge.

To measure the E.M.F. across the tuned circuit, the E.M.F. is amplified a known amount (using a special four-stage amplifier) and is rectified so that its R.M.S. value may be determined with a direct current instrument; it is important that the rectifier be used only within the range in which it gives a current proportional to the square of the impressed voltage. The tuned circuit picks out a narrow band of frequencies present in the fluctuations and, since the amplification is not the same for frequencies slightly different from the resonance frequency, correction is made for this by a factor F, for which a mathematical expression is derived, and values are obtained by integration or summation. Calibration of the amplifier is avoided by substituting, after each reading of the amplified and rectified "Schrot" current, a measured pure sine voltage across the terminals of the tuned circuit, which is adjusted to give the same rectified current as the effect. Calling this v_1, the actual mean square shot voltage,

$$v_0^2 = v_1^2/\text{F. and } I^2 = Cv_1^2/LF,$$

where L is the inductance of the tuned circuit. It was found that with currents limited by the space charge, the observed effect might be only

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20 per cent. of the theoretical value, but with currents limited only by the temperature of the filament, the agreement was within 1 per cent. The phenomenon may therefore be used to study the effect of the space charge on the electronic evaporation.

5. **Determination of Electronic Charge from Measurements of Shot-effect in Aperiodic Circuits :** N. H. WILLIAMS and H. B. VINCENT : *Phys. Rev.*, 1926, **28,** 1250-64.

An outstanding difficulty in the measurement of the electronic charge from the " Schrot " effect in a tuned circuit has been the retro-action into this circuit. Elaborate screening devices have hitherto been adopted. Former methods of dealing with this difficulty have added tedious computations to Schottky's original calculation. The solution presented here yields the result in such a form that much of this former labour is eliminated ; the solution is of great mathematical simplicity, and the measurement of the radio-frequency resistance and resonance frequency becomes unnecessary. The treatment may be generalised to include the non-resonant case if the assumption be made that the resonance curve of the amplifier is symmetrical. A non-inductive resistance may be used as the medium of coupling, and experimental work with this system has given a mean value for the electronic charge of $4 \cdot 774 \times 10^{-10}$ e.s.u. In work under conditions other than those of temperature limitation, the diminution of fluctuations seems to be due to high current densities rather than to the space charge in the region of the filament.

6. **Some Extensions of Theory and Measurement of Shot-effect in Periodic Circuits :** H. B. VINCENT : *Proc. Nat. Acad. Sci. Wash.*, 1927, **13,** 774-85.

Some additions are made to the method discussed in the previous paper (see Abs. 4.1.2/5) for computing the effect in a periodic circuit of the probability fluctuations of the anode current in a vacuum tube. A mechanical integrator is described which greatly improves the accuracy of the method. The measurements that were made lead to values of the electronic charge of the order of $1 \cdot 70 \times 10^{-19}$ coulomb, which is somewhat greater than the accepted value ($1 \cdot 59 \times 10^{-19}$), although the total error is estimated at less than 2 per cent. Measurements of the " Schrot " effect at different anode currents showed a slight diminution in the ratio of observed fluctuation to calculated fluctuation as the anode current increased fr om 1 to 5 milliamperes

4.1.3. INTER-ELECTRODE CAPACITIES AND INPUT IMPEDANCE

1. **Dependence of the Input Impedance of a Three-electrode Vacuum Tube upon the Load in the Plate Circuit :** J. M. MILLER : *Sci. Pap. Bur. Stand. Wash.*, 1919, **15,** No. 351, 367-85.

The experimental portion of this paper gives the results of measurements upon the input impedance of particular valves.

Close agreement with theoretical results is obtained at medium frequencies only.

(See also Abs. 1.2.2/10.)

2. **Note on the Input Impedance of Vacuum Tubes at Radio Frequencies :** J. WEINBERGER : *Proc. Inst. Radio Eng.*, 1920, **8,** 334-9, 340-1.

Experiments are described which show that the input impedance of a valve at radio frequencies is dependent upon the nature of the anode circuit load. The reduction in the effective resistance of the input circuit is noted for the case of an inductive load, and pure resistance and capacity resistance combinations are briefly discussed.

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The experimental method has been severely criticised by L. M. HULL and the results obtained are very approximate. However, the paper anticipates MILLER'S work.

3. **A Method of Measuring very small Capacities :** L. HARTSHORN : *Proc. Phys. Soc. Lond.*, 1924, **36**, 399-404.

The Schering modification of the Wien bridge is described and is shown to be suitable for the measurement of very small capacities (of the order of 1 micro-micro-farad) at telephonic frequencies. Means of securing (1) great sensitivity ; (2) fineness of adjustment ; and (3) the elimination of capacities to earth and capacities due to the presence of connecting leads, are indicated, so that it is possible to measure the true capacity between any two conductors, e.g., the electrodes of a thermionic valve. As a test of the method, it is used to measure the capacity between two steel balls of 1 cm. radius and at distances apart of 2·25 cm. to 8 cm.

4. **The Inter-electrode Capacities of Thermionic Valves :** L. HARTSHORN and T. I. JONES : *Experimental Wireless*, 1925, **2**, 263-73.

The Schering bridge adopted in previous work is here used to examine the inter-electrode capacities of various types of valves. The capacities in the R type of valve are much larger than in the V 24 type, due to the leads in the " pinch " of the valve in the former case. It is shown that the holder contributes largely to the value of the capacity and the actual capacity of different types of holder is measured.

The upper limit of the frequency of a valve oscillation generator is governed by the valve capacity, and this limit is calculated for various types of oscillation circuit.

Finally, it is shown that the capacity is always impure, being associated with a conductance. Power factor measurements have been taken to ascertain the value of these conductances. Finally, the effects of various types of valve base are discussed.

5. **Performance of Amplifiers :** H. A. THOMAS : *J. Inst. Elect. Eng.*, 1926, **64**, 253-68, 268-78 ; *Experimental Wireless*, 1926, **3**, 15-18 ; *Elect. Rev. Lond.*, 1926, **98**, 73-74.

(See Abs. 1.2.7/4, 1.4.1/16, 3.4/9, 4.2/8, 4.4.1/7, 4.5/7, 4.6/1.)

6. **The Input Impedances of Thermionic Valves at Low Frequencies:** L. HARTSHORN : *Proc. Phys. Soc. Lond.*, 1927, **39**, 108-23.

(See Abs. 1.2.7/9.)

7. **Measurements on Radio-frequency Amplifiers :** R. L. SMITH-ROSE and H. A. THOMAS : *Wireless Wld.*, 1927, **20**, 131-5, 171-7, 199-202, 237-9.

(See Abs. 4.2/12.)

8. **Measurement of Vacuum Tube Capacities by a Transformer Balance :** H. A. WHEELER : *Proc. Inst. Radio Eng.*, 1928, **16**, 476-81.

A complete portable equipment is described for the measurement of the direct capacities of valves in laboratory or factory testing. The valve capacity is compared with a standard variable condenser by means of a transformer-balance circuit, whose balance is independent of the frequency (about 1,500 kilo-cycles per sec. being used). The method consists in adjusting the standard condenser acting as the " Neutrodyne " condenser until balance is obtained when the valve is supplied with a radio-frequency input. Designs are proposed for the standard condenser and the transformer, and suggestions are made for the further improvement of this equipment. The advantage

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of such an arrangement is that the effective capacity is measured at the actual radio frequency for which the receiver is designed to operate.

9. **A Bridge Method for Measurement of Inter-electrode Admittance in Vacuum Tubes :** E. T. HOCH : *Proc. Inst. Radio Eng.*, 1928, **16,** 487.

A description is given of the Colpitts-Campbell bridge as applied specifically to the measurement of direct admittances in valves. The measurements were obtained at a frequency of 1,000 p.p.s., and the difficulties encountered with this bridge at radio frequencies are discussed. The capacity and conductance of several American valves is given.

10. **A Direct-capacity Bridge for Vacuum-tube Measurements :** L. WALSH : *Proc. Inst. Radio Eng.*, 1928, **16,** 482-6.

A direct-capacity bridge is described which permits the measurement, at a single setting, of a capacity associated with other capacities in a system having more than two terminals, such as the grid-anode capacity of a valve. The usual methods for determining the grid-anode capacity of a valve involve measurements of the various combined inter-electrode capacities and the deduction therefrom of the desired capacity. The author takes a standard form of capacity bridge and connects the grid and anode to the two terminals provided for the unknown capacity. The filament is then connected to the common point of the two ratio arms, which is earthed. The filament-anode capacity is thus effectively inserted across the detector, and does not affect the balance. The filament-grid capacity is in parallel with one of the ratio arms, and it is shown that its effect can be balanced by a resistance inserted in series with the standard capacity.

11. **Inter-electrode Capacities and Resistance Amplification :** L. HARTSHORN : *Experimental Wireless,* 1928, **5,** 419-30.

(See Abs. 3.2.2/26.)

12. **A Method of Treating Resistance-stabilised Radio-frequency Amplifying Circuits :** B. L. SNAVELY and J. S. WEBB : *Proc. Inst. Radio Eng.*, 1929, **17,** 118-26.

(See Abs. 1.4.1/24.)

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4.2. RADIO-FREQUENCY AMPLIFIERS

THE behaviour of a radio-frequency amplifier can be expressed in terms of three characteristics :—

(1) The voltage amplification, expressed as the ratio of the radio-frequency voltage supplied to the rectifier to the voltage applied to the grid of the first stage.

(2) The input impedance of the system, expressed as an equivalent load which acts as a shunt to the tuned input circuit. Such a load varies the effective resistance of this tuned circuit.

(3) The wave form distortion due to the amplification process, expressed as the magnitude of the harmonics produced by the passage of the signal through the various stages of the amplifier.

Both the voltage amplification and the input impedance are dependent upon the type of rectifier employed and upon its equivalent load, and this equivalent load will in turn be dependent upon the nature of the audio-frequency amplifier and the output mechanism. Consequently the voltage amplification and the input impedance can only be expressed in terms of specified output conditions. Again, these characteristics are dependent upon the frequency, and so it appears that the measurements required to specify the performance of any radio-frequency amplifier are :—

(1) A voltage amplification-frequency characteristic for the working range of frequency in which the normal rectifier load and the output circuit arrangements are not disturbed by the measuring arrangement.

(2) An input impedance-frequency characteristic obtained for the same conditions. This can be expressed in several ways ; possibly the best method is to represent the impedance in terms of a shunt condenser and a shunt resistance, both varying with frequency.

(3) A determination of the harmonics produced by the amplification process. This latter property affects the behaviour of the rectifier and in consequence the overall performance of the complete receiver.

In any measurement of the properties of a radio-frequency amplifier, various important points must be carefully borne in mind if the measurement is to be such as truly to represent the actual working conditions.

(1) The measuring apparatus must not modify the normal behaviour of the amplifier, i.e., no instruments must be inserted which modify the general circuit arrangements.

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(2) The circuit which supplies the input terminals of the amplifier must have a low impedance, otherwise retro-action may occur and such retro-action will modify the amplification factor.
(3) The instrument used to measure the output must be of such a nature that it inserts a negligible load.
(4) The input and output voltages existing during any test must be of the same order as those encountered in actual operation.
(5) The earthing or screening arrangements must be the same in the test arrangement as in actual operation.
(6) The percentage of modulation adopted in the tests must conform to the normal usages of the amplifier.

An examination of the methods that have been adopted to measure amplification at radio-frequencies reveals the manner in which the above conditions have only partially been satisfied.

The method of measurement used by SMITH and NAPIER (see Abs. 4.2/1) made use of a telephone balance, by which means the output was kept constant under two different conditions (*a*) when the input was applied directly to the grid of the stage including the telephone, and (*b*) when the input was applied to the grid of a stage preceding the output stage. The ratio of the two E.M.F's. gave the amplification of the stage under consideration.

The accuracy of measurement was limited by the sensitivity of the human ear to small changes of sound amplitude. This is given as 5 per cent. by some experimenters and 30 per cent. by others; about 10 per cent. is all that can be expected. The modulation was obtained by means of a mechanical chopper which varied the radio-frequency wave-length, and the rectangular form of modulation so produced is liable to introduce transient effects of a serious nature. The range of amplification measurement possible is also limited.

PIRANI (see Abs. 4.2/2) used a capacity potentiometer method, in which the audio-frequency output was balanced against a definite output from the modulator, this output acting as a comparison standard. The measurements do not give the actual values of the input or output voltages or the modulation percentage, and are useful only for approximate determinations. The capacity of the input source may produce retro-active phenomena, which would not exist if a low impedance source was used, but the effect of this source is not so serious as that of a tuned input circuit.

The method due to JORDAN (see Abs. 4.2/3) gives an amplification measurement by means of a telephone comparison without an actual measurement of the values of the input and output voltages. The method involves the calibration of a mutual inductance at radio-frequencies, and this introduces a source of error, owing to the difficulty of making such inductance measurements.

BLEY (see Abs. 4.2/4) dispenses with the modulated source and the need for rectification, a valve voltmeter being used to measure

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the output voltage. A comparison is made of the mutual inductance between two coils required to give a certain valve-voltmeter output when this voltmeter is connected directly to the secondary circuit of the mutual inductance and when the amplifier is inserted between the mutual inductance and the voltmeter.

The valve voltmeter throws a very uncertain load upon the mutual inductance, and the difficulty of accurately calibrating the inductance is still present. However, the necessity for adequate screening is realised and an attempt is made to minimise the effects of magnetic induction.

In all these methods, retro-active effects are not excluded, due to the relatively large impedance of the input system, and the use of mutual inductances limits the possible frequency range of the measurements. FRIIS and JENSEN (see Abs. 4.2/6) pointed out the necessity for the use of a low impedance input circuit, and in their method employ a low resistance potentiometer fed from a modulated radio-frequency source. The output was measured by a valve voltmeter. The maximum amplification which they could measure was 500.

The method developed by THOMAS at the National Physical Laboratory in 1925 (see Abs. 4.2/8) used a low resistance potentiometer and a vibration galvanometer to record the output. By tuning this instrument, very weak telephonic outputs could be used, and the method is still useful for the calibration of amplifiers where telephone signals only are required. For loud speaker outputs, the valve voltmeter is preferable, though care must be taken in the manner of its connection. The screening of the various parts of the system from stray fields is most important, and the oscillators and potentiometers are elaborately screened ; the amplifier under test is screened from external interference by being situated in a specially shielded room.

For very high amplification ratios, difficulty is experienced in the design of the potentiometer, since, at high radio-frequencies the inductance of the potentiometer wire gives a greater reactance than the resistance of the wire. In the measurements made by HULL (see Abs. 4.2/9) on his screen-grid valves, amplifications as high as 2,000,000 were measured, and for this purpose the voltage drop across a short length of copper wire was used to supply the input to the amplifier. The inductance of such a wire can be calculated and, if a concentric return is used, the system is also effectively screened. OAKLEY (see Abs. 4.6/6) employs the same method in his " attenuator." This arrangement is satisfactory for frequencies up to 1,000 kilo-cycles per sec., and if care is taken, it can be used for frequencies as high as 30,000 kilocycles per sec.

DYE (see Abs. 4.2/7) uses a current transformer in conjunction with a potentiometer to obtain small known E.M.F's. at radio-frequencies. By suitable design of the iron path, the ratio is

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accurate at such frequencies, provided the resistance load across the secondary obeys certain conditions. A current reduction of 100 to 1 is easily obtained by this instrument, and by suitable turn ratios, a large range of voltage can be given with a limited primary current range.

The determination of the input impedance of a multi-stage radio-frequency amplifier is more difficult than the measurement of its overall amplification. The input impedance can be expressed in terms of a condenser and a resistance in parallel, and usually it is the value of the resistance component which is most required, since this modifies the effective resistance of the input tuned system and consequently the overall performance.

Three methods of solution are possible :—

(i) A bridge measurement. In this case it would be necessary to determine the impedance of a non-inductive resistance of moderately low value when shunted by the amplifier, from which knowledge the effective input impedance could be calculated. However, the amplifier must be supplied with a very small voltage of the same order as the normal working input voltage and this means an exceedingly small voltage across the bridge. The difficulties of using a bridge at high radio-frequencies are many, and it appears that the utility of this method of measurement is severely limited by the several conditions which need to be fulfilled in any measurement on a multi-stage amplifier.

(ii) An inserted resistance method. This is a well-known method of determining the effective radio-frequency resistance of a tuned system. By inserting a non-inductive resistance into the tuned circuit, the drop in the measured output gives a measure of the radio-frequency resistance. If the resistance of the tuned circuit without the amplifier is measured in any of the usual ways, the difference in resistance produced by the amplifier will give the effective resistance component of the input impedance. The capacity component can be obtained by noting the different tuning condenser values for a definite frequency with and without the amplifier connected.

This method is suitable for long wave lengths, but at short wave lengths the alteration in the mechanical configuration of the tuned circuit, due to the presence of the inserted resistance, may be serious.

(iii) A resonance curve method. In this case, resonance curves
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the two components of the input impedance. This method has been found satisfactory at medium radio-frequencies, since the measurements can be obtained without disturbing the amplifier or the tuned circuit in any manner.

No published information can be found and there is a need for an examination of the three methods with a view to ascertaining which is most suitable.

The wave form distortion produced by a radio-frequency amplifier will affect the overall performance, since the operation of the rectifier will be modified by the presence of harmonics in the output wave form from the amplifier. No work has been done on this subject, and here again there is room for a study of this property of the radio-frequency amplifier.

The tendency in the progress of apparatus suitable for high radio-frequency measurements has been towards the improvement of methods which supply an input as nearly identical as possible in character to the actual input during operation, and towards the use of methods of measuring the output which introduce the smallest possible disturbing effect upon the amplifier. The first consideration requires a modulated radio-frequency source and perfect screening, together with accurate means of measuring very small voltages. It is now realised that, since retro-action plays such an important part in the behaviour of a radio-frequency amplifier, computations of the amplification characteristic based upon the measured performance of successive stages are worthless.

The complexity of the effects which exist in a multi-stage radio-frequency amplifier is such that a separation of the overall amplification into its constituent parts is almost impossible except perhaps at the lowest radio-frequencies. The unknown quantities, even in a completely screened amplifier, are numerous, and consequently a complete theoretical examination of the problem of amplification is not possible except in the most general manner. Experimental work has not therefore been directed towards the confirmation of any theory, but has yielded specific results which show, in a general manner, the major properties possessed by the particular amplifiers tested. The main desire of experimenters has been to evolve a satisfactory method of measurement, and only in certain specialised cases has this been done.

A few general results have been obtained, and may be stated as follows :—

(1) The amplification of a two or three-stage amplifier at low radio-frequencies can be calculated approximately from the known constants of the valves and circuits.

(2) At medium radio-frequencies, the amplification per stage appears to be much lower than that which would be expected from a theoretical analysis. (TRENDELENBURG

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and RIEGGER (see Abs. 4, 2/5) give experimental values of 4·8 and 6·1 for two cases, where the theoretical value was about 7 and 20 respectively, and the author has obtained values of 4 where the theoretical value was about 8.)

(3) The overall amplification of a multi-stage amplifier is rarely found to be equal to the product of the measured values of the individual stages.

(4) The retro-active effects contribute so largely to the overall performance that the characteristic properties of the input tuned circuit (viz., its resistance and magnification factor), and of the tuned anode circuits in the radio-frequency stages are often completely eclipsed.

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1. **On the Measurement of Amplification given by Triode Valves at Audible and at Radio-frequencies :** F. E. SMITH and H. G. NAPIER : *Proc. Phys. Soc. Lond.*, 1920, **32,** 116-32.

The section of this paper dealing with radio-frequency amplification describes a method in which a radio-frequency source is modulated by a mechanical chopper at a frequency of 800 p.p.s. The rotating switch inserts an inductance as a shunt across the tuning condenser, giving an oscillation at a different wave length. A given output is obtained when the input is applied directly to the rectifier and then again when a weaker input is applied to the tuned input circuit of the amplifier, the output being maintained the same as before. This change is effected suddenly by means of a switch, and a telephone balance is obtained.

Apart from the difficulties of using tuned circuits and the ambiguity of the results obtained by their use, an additional objection occurs in this arrangement due to the approximately square wave form modulation. It is possible that serious harmonics may be produced and that the shock impulses given to the amplifier will produce transient phenomena of a complex nature.

(See also Abs. 4.4.1/1.)

2. **Amplification Ratio of Radio-frequency and Audio-frequency Amplifiers :** M. PIRANI : *Jahrb. drahtl. Telegr.*, 1920, **16,** 2-30.

The latter half of this paper deals with measurements at radio-frequencies. A modulated radio-frequency oscillator supplies a capacity potentiometer by which means a small E.M.F. can be applied to the amplifier. The audio-frequency output is balanced against a definite output from the modulator, which acts as a comparison standard and is not calibrated in volts.

The measurement gives no absolute values of percentage modulation, input or output voltages, and the effect of the capacity load due to the potentiometer system may produce important retro-action effects, although the effect of a capacity potentiometer is not likely to be so serious as that of a tuned input circuit.

(See also Abs. 4.4.1/2.)

3. **A Method of Measuring the Amplification of a Radio-frequency Amplifier :** F. W. JORDAN : *Proc. Phys. Soc. Lond.*, 1920, **32,** 105-15.

The method of measurement consists essentially in balancing the telephone output given by a modulated radio-frequency source against a known signal directly applied by means of a potentiometer to the telephone.

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The audio-frequency oscillator consists of a valve maintained tuning fork which provides a constant frequency source of pure sinusoidal E.M.F. of a few volts. This output is inserted in series with the radio-frequency oscillator high tension battery lead and thus gives a known modulated output. By means of an intermediate tuned circuit, the output is led to a mutual inductance coil which inserts the radio-frequency E.M.F. into the tuned input circuit of the amplifier under test. The output is balanced against a standard audio-frequency E.M.F. derived by a potentiometer fed from the original modulator.

In making measurements of amplification, the E.M.F. existing across the tuned circuit condenser is applied to the rectifier only, and the output is adjusted to correspond to that of the standard signal. The amplifier is now inserted and a much weaker input is applied, the output being the same as before. The ratio of the mutual inductances in the two cases gives the amplification.

In considering the application of this method for really accurate measurement work, the following points must be noted. In the first place, the limiting accuracy with which any telephone balance can be obtained is of the order of 5 per cent., since the human ear is very insensitive to small changes of intensity. Secondly, the accurate determination of the mutual inductance at radio-frequencies is not a simple matter, and the difficulties are increased if the range of calibration has to be very large in order to measure large amplification factors. Thirdly, the presence of a tuned input circuit will seriously modify the behaviour of the amplifier, and the measured result obtained cannot be regarded so much as a property of the amplifier alone as of the tuned circuit and amplifier together, and in this latter case, the tuned circuit is seriously modified by the inducing coils. Amplification measurements must use a very low impedance input circuit if retro-action effects are to be eliminated from the measurements.

No actual knowledge of the value of the input and output voltages and the modulation percentage is obtained by this method, and the amplification figure could not be obtained if a specification was given of the nature of the input voltage with which the amplifier was to be supplied.

4. **Experimental Investigation of Radio-frequency Amplifiers :** A. BLEY : *Arch. Elektrotech.*, 1923, **12**, 124-43.

The voltage amplification of several valves is measured with both resistance and inductance coils in their anode circuits at frequencies from about 300 to 12 kilo-cycles per sec. (1,000 to 24,000 metres), and compared with the theoretical amplification derived from the static characteristics of the valves, regard being paid to the self-capacities of the resistances and coils. In this method of measurement, the modulator is dispensed with, pure continuous waves being applied by means of an oscillator via a linked circuit and a mutual inductance coupling. The output is measured by a valve voltmeter ; the direct anode current is balanced out by means of a primary cell and a potentiometer in the usual manner ; such an instrument gives deflections of the galvanometer which are approximately proportional to the square of the applied E.M.F. which can be calibrated at moderately low frequencies by means of a potentiometer or a calibrated mutual inductance.

The valve voltmeter is first connected across the input terminals of the amplifier, and the mutual inductance is adjusted so as to give a readable deflection of the voltmeter. With a weaker coupling, the same deflection is obtained with the voltmeter connected to the output terminals of the

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amplifier. The ratio of the two values of the mutual inductance gives the amplification.

Its disadvantages are, however :—

(i) Possible retro-action effects which may be produced due to an inductive input circuit.

(ii) The loading effect of the valve voltmeter upon the previous stage circuit is serious when used in the manner described by this writer.

(iii) The difficulty of calibrating a large ratio mutual inductance at radio-frequencies, although it appears from Bley's results that he used the method for amplification factors up to 1,200.

The method is superior to the earlier methods, and the necessity for careful screening is appreciated.

5. **Amplification of Short Waves :** F. TRENDELENBURG and H. RIEGGER : *Z. tech. Phys.*, 1924, **5**, 495-500.

This paper deals with a method of measuring amplification and gives interesting results obtained on one, two, three and four stages of a choke-capacity-coupled amplifier working between 130 and 500 metres.

The method of measurement consists in supplying a known voltage by means of a step-down mutual inductance in which the coils are of the toroidal form. Both the oscillator and the amplifier are heavily screened. The results given in graphical form give a clear picture of the amplification obtained from a choke-capacity-coupled amplifier. The first set of curves give the performance of such an amplifier working over a range of 1·0 to 1·66 kilo-cycles per sec. (300 to 500 metres). It is interesting to note that the maximum amplification per stage was 4·8 with a valve having a voltage factor of 9. The second set of curves gives a performance of a second amplifier working over a range of 2·0 to 2·3 kilo-cycles per sec. (150 to 130 metres). Here the maximum amplification obtained for one valve was 6·1 while its voltage factor was 25.

It is clear that in this latter case, the overall factor falls much below the product of the factors of the individual stages ; this effect is not nearly so noticeable on the 300 to 500 metre wave-length band. These results are in accordance with the most recent measurements and support the view that the voltage amplification of a radio-frequency stage is smaller than the value obtained by a simple theoretical prediction.

6. **High Frequency Amplifiers :** H. T. FRIIS and A. G. JENSEN : *Bell System Tech. J.*, 1924, **3**, 181-205.

The latter part of the paper deals with a method of measuring the performance of radio-frequency amplifiers. The authors point out the necessity for using a low resistance potentiometer to give the required known input E.M.F., the maximum resistance suggested being 100 ohms, above which value the authors assert that the potentiometer introduces retro-action and so produces an error. The method of measurement consists in applying a modulated radio-frequency current through a known resistance potentiometer, the current being measured by a vacuum-junction ammeter. The input terminals of the amplifier are connected across this potentiometer, and the output is measured by means of a valve voltmeter. The maximum amplification which can be measured is 500.

To obtain the overall amplification of a system comprising an amplifier and a tuned circuit, the authors use a valve voltmeter connected across the tuned circuit, and by inducing an E.M.F. into this tuned circuit, the voltage across the tuning condenser is obtained with and without the amplifier attached. If the voltage is less when the amplifier is present, the

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net effect of the amplifier is that of a positive shunt load; if, however, the voltage across the tuning condenser is greater when the amplifier is present, retro-action is taking place.

The assumptions made in these measurements are two-fold. Firstly, that the valve voltmeter throws no load on the tuned circuit and in no way affects the inter-action between the amplifier and its tuned circuit; and secondly, that the retro-action effects are independent of the amplitude of the potential applied to the amplifier. This latter assumption is undoubtedly unjustifiable, since, to obtain a readable deflection on the valve voltmeter, the input voltage must be of the order of 0·5 volt; the amplifier will obviously be saturated with such an input and its normal operating characteristics destroyed. It is possible that the former assumption is valid under certain conditions of the valve voltmeter, the tuned circuit, and the frequency.

(See also Abs. 1.2.3/5.)

7. **Current-transformer Methods of Producing Small Radio-frequency Voltages and Currents**: D. W. DYE: *J. Instn. Elect. Eng.*, 1925, **63**, 597-608.

The author outlines a method of producing known small radio-frequency voltages. The experimental arrangement consists of a current transformer, across the secondary of which is a known resistance, a thermo-junction being used to measure the primary current. Voltages from a few microvolts to a few tenths of a volt are conveniently produced. The nature and the extent of the errors of the method are discussed, and the precautions necessary to reduce these to a negligible amount indicated. The applications of the method to produce a very small current or voltage in an inductive coil, and to calibrate an amplifier or a receiving system are given, and an appendix gives a simple analytical treatment of the iron losses of a current transformer at radio-frequencies.

8. **The Performance of Amplifiers**: H. A. THOMAS: *J. Instn. Elect. Eng.*, 1926, **64**, 253-68, Disc. 268-78.

Section I of this paper deals with the researches which have been carried out at the National Physical Laboratory for the RADIO RESEARCH BOARD. A radio-frequency oscillator is used as the source supplying a low resistance potentiometer. The radio-frequency current is modulated by means of a 1,000 cycle oscillator, and the modulation percentage can be determined by the use of an electrostatic voltmeter. The output operates a vibration galvanometer tuned to the modulation frequency; this output stage is calibrated directly from the modulator. The leads and oscillators are carefully screened, and details are given of the earthing devices adopted. The section concludes by presenting results of measurements on several amplifiers in graphical form.

(See also Abs. 1.2.7/6, 1.4.1/16, 3.4/9, 4.4.1/7, 4.5/7, 4.6/1.)

9. **Measurements of High Frequency Amplification with Shielded-grid Pliotrons**: A. W. HULL: *Phys. Rev.*, 1926, **27**, 439-54.

Tests are described on cascade radio-frequency amplifiers using specially constructed valves in which the control grids are electrically screened from the anodes. These valves are almost free from internal retro-action, so that a large number may be operated in cascade.

The minute input voltages were obtained from the calculated inductive drop in a short section of straight copper rod surrounded by a concentric return conductor. It is stated that this method is reliable for frequencies up to 1,000 kilo-cycles per sec. (300 metres); at 10,000 kilo-cycles per sec.

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(30 metres) care is required in connecting the input system to the amplifier, to avoid capacity shunting effects.

The voltage amplification per stage depends upon the anode-circuit impedance, and was found to be 200 at a frequency of 50 kilo-cycles per sec. (6,000 metres), 40 at a frequency of 1,000 kilo-cycles per sec. (300 metres), and 7 at a frequency of 10,000 kilo-cycles per sec. (30 metres). Total voltage amplifications as high as 2,000,000 were obtained at a frequency of 1,000 kilo-cycles per sec. (300 metres). It is shown that this is the largest amplification that can be usefully employed at any frequency, due to "Schrot" effect. At a frequency of 10,000 kilo-cycles per sec. (30 metres) a total amplification of 10,000 was obtained from four stages. Examples are given, showing some applications of such an amplifier, e.g., the amplification of small photo-electric currents.

10. **The Performance of an Intermediate-frequency Amplifier :** M. G. SCROGGIE : *J. Instn. Elect. Eng.*, 1927, **65**, 644-7.

Measurements were carried out on the intermediate-frequency amplifier of a supersonic-heterodyne receiver designed primarily for broadcast reception. The method used consisted in feeding the input with an alternating E.M.F. of known amplitude and frequency, and measuring the output by a valve voltmeter. The disturbing effect of a measuring instrument for indicating the output voltages is avoided by making the last valve itself serve as the voltmeter, thus preserving working conditions. Three representative types of valves were used in turn, and it is shown that, for a prescribed standard of cut-off at the extremes of the side-bands, the amplification obtainable increases with the mutual conductance of the valves. The best type of valve to use in various circumstances is deduced from the results obtained.

11. **Measurements of Radio-frequency Amplification :** S. HARRIS : *Proc. Inst. Radio Eng.*, 1927, **15**, 641-8.

The author describes a method of measurement in which the input is kept constant at some definite but unknown value. The output circuit is connected to a 20,000 ohm potentiometer, a known tapping from which is taken to a valve voltmeter. Any reading on the valve voltmeter is taken with a given setting of the step-down potentiometer. One stage of the radio-frequency amplifier is now removed and the potentiometer ratio is increased until the output from the valve voltmeter is the same as before. The ratio of the two settings of the potentiometer gives the amplification of the stage which was removed. By repetition of this process for each stage, the total amplification is given as the product of the successive results.

It is obvious that this method is only applicable to cases where retro-action and loading effects are negligible ; in fact, the author states that the method is applicable only to screened non-retro-active amplifiers. If the method is so limited, it appears that nothing is gained on the simpler method of measuring the individual stage factors and taking their product, since such a procedure is equally justifiable for a non-retro-active amplifier.

12. **Measurements on Radio-frequency Amplifiers :** R. L. SMITH-ROSE and H. A. THOMAS : *Wireless Wld.*, 1927, **20**, 131-5, 171-7, 199-202-237-9.

In the first section of this paper, a review is given of the different methods of measuring the amplification of radio-frequency amplifiers. In Section II, the improved National Physical Laboratory method is described. The use of careful screening arrangements was found to be necessary, and a full description of these arrangements is given, together with some interesting characteristics of various types of amplifiers, shown in graphical form. In Section III, a discussion of the effects of the input impedance of the amplifier

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upon the tuned circuit to which it is connected are discussed in a general manner, after which the method of measurement is described. It was found inadmissible to employ the inserted resistance method since retro-action phenomena are modified by its presence. The method finally adopted consists in obtaining the resonance curve of the tuned circuit by slightly varying the frequency of a radio-frequency oscillator inductively coupled to it. By comparing this resonance curve with that obtained for the tuned circuit alone, the change in effective resistance due to the presence of the amplifier can easily be calculated. Results are given showing the manner in which the amplifier may behave as a positive or negative load on the tuned circuit. The fourth section deals theoretically with the properties of the intermediate-frequency amplifier of a supersonic-heterodyne receiver, showing the necessary selectivity required for the elimination of local interference.

13. **Inductive Coupling with Primary and Secondary Tuning Connecting Valves :** H. KAFKA : *Jahrb. drahtl. Telegr.*, 1927, **30**, 44-52.

(See Abs. 1.3/12.)

14. **On the Distortionless Reception of a Modulated Wave and its Relation to Selectivity :** F. K. VREELAND : *Proc. Inst. Radio Eng.* 1928, **16**, 255-80, Disc. 494-7, 671-3, 848-50.

(See Abs. 1.2.1/24.)

15. **Method and Apparatus for Investigating High-frequency Amplification :** M. v. ARDENNE : *Elektrotech. Z.*, 1928, **49**, 1675-8.

This paper describes in detail an apparatus for measuring directly the amplification of complete radio-frequency amplifiers. A local oscillator serves as a source and is inductively coupled to the grid circuit of the first amplifier. An electrometer serves to measure the grid volts applied to this valve. The output voltage variation of the last valve of the amplifier is measured by means of a valve voltmeter. Curves are given showing the radio-frequency amplification of Loewe valves and a particular screen-grid valve. It is claimed that the method is accurate and that the readings can be obtained quickly.

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4.3.—THE PERFORMANCE OF RECTIFIERS

Much of the experimental work hitherto carried out on the performance of rectifiers suffers from a lack of systematic analysis, with a consequent uneconomic design of experiment. With a few notable exceptions, the published papers record work on particular cases and do not lead to general deductions applicable to design.

In any practical application of a rectifier, there is a frequency transformation of electrical energy. Three distinct elements are involved in the process: an input circuit; the rectifier itself with its specific characteristics; and, finally, the output circuit or rectifier load. In general there will be certain interactions between these elements. Thus the input impedance of the rectifier will react on the input circuit associated with it. In many cases, the specific characteristics of the rectifier will be affected by the output circuit load. It should be possible, however, by suitably designed experiments to isolate the specific characteristics of the rectifier and to determine its mutual relationships with the load in such a way as to permit of the approximate predetermination of its performance under any reasonably assignable conditions, and, further, to specify the most efficient or, where efficiency is not the important consideration, the most suitable conditions of operation. It is not too much to say that no paper which has hitherto appeared satisfies the above requirements in full for any form of signal rectification.

An early attempt at a systematic description of rectifier characteristics was made by COLEBROOK in 1924 (see Abs. 4.3/8). The scheme of specification proposed by him is described more fully in the section dealing with the subject of rectification in general. The paper referred to records experimental determinations of the effective rectified voltage and internal resistance in the simple case of the crystal rectification of pure continuous waves. Theoretical determinations of the corresponding quantities in relation to the rectification of modulated radio-frequency potential differences are also given but no measurements were made. The input characteristics of such rectifiers are less satisfactorily treated and no measurements were attempted.

A further paper by the same author extended the same treatment to valve rectification and contains a quantity of useful information on the specific characteristics of small valves in grid and anode-circuit rectification. Here again, however, the measurements were confined to the simple case of pure continuous waves, over rather unduly restricted ranges of voltage. Further, the treatment of input impedance is unsatisfactory, being confined to that part of the input impedance which is due to the rectification process itself,

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ignoring the usually more important component which is due to the unavoidable inter-electrode capacity effects of the valve.

A considerable volume of work has been published recording measurements of the rectification of pure continuous waves. Such measurements are comparatively simple, except in the matter of input impedance, which factor is usually left out of account, probably because of the difficulty of the measurements. In most cases the input potential differences are obtained by means of measured currents in known resistances. In one case (see Abs. 4.3/6) a condenser potential-divider is used, a convenient method which, however, would probably not be applicable in a simple form at high radio-frequencies.

Rather less work has been carried out on the rectification of modulated or heterodyned continuous waves. The chief difficulties in this work are the production of a known modulated input and the selective measurement of the single modulation frequency component in the somewhat complex output from the rectifier. CHAFFEE and BROWNING (see Abs. 4.3/14) used a bridge method in which the modulation frequency output was balanced against a current derived from the modulator. They made use of the conception of effective modulation frequency, rectified voltage and internal resistance both in grid and anode-circuit rectification, but did not give any detailed applications of these ideas to design. In another case the modulation-frequency output was compared audibly against a known modulation-frequency current. This is a simpler method, but takes no account of wave form differences which might invalidate the audible comparison. P. DAVID (see Abs. 4.3/20) used a valve voltmeter for the measurement of the modulation-frequency output, this being capacity-resistance coupled to the rectifier, with the resistance short-circuited for the radio-frequency components by a tuned "acceptor" circuit. This method is open to some criticism in respect of audio-frequency wave-form, but DAVID does not claim a high degree of accuracy. It was sufficiently accurate for the purpose he had in view. KUHLMNANN (see Abs. 4.3/9) used a tuned vibration galvanometer. Given a suitable and simply controlled instrument of this type, the method would appear to be very well adapted to its purpose, since it would respond only to the desired frequency component. MEDLAM and OSCHWALD (see Abs. 4.3/21) used an audio-frequency filter arrangement for the isolation of the desired frequency in the output circuit. An alternative method which does not appear to have been utilised so far is to employ a very selective audio-frequency amplifier in conjunction with some suitable alternating current measuring instrument (such as a thermo-junction). Such amplifiers can be given almost any desired selectivity by means of retro-active control of the tuned circuit damping, or by the use of circuits of very low decrement.

With regard to heterodyne rectification, no completely satisfactory experimental work applicable to present day conditions has

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yet been published. The original paper by ARMSTRONG (see Abs. 4.3/1) would almost certainly require revision in view of the improvement both in receiving valves and methods of measurement. The theoretical conclusions arrived at by APPLETON and TAYLOR (see Abs. 2.3.2/14) with regard to optimum local oscillator amplitude are supported by indirect experimental evidence, measurements of the actual beat frequency output not being given. The most important recent contribution to the subject is that made by P. DAVID (see Abs. 4.3/20), who shows that in long-wave heterodyne rectification and in supersonic-heterodyne rectification the greatest output is obtained when through the agency of grid-circuit rectification of the local oscillator E.M.F., a mean grid potential appropriate to anode-circuit rectification is arrived at. That is to say, although the receiving circuit may be connected as for grid rectification, it is invariably anode-circuit rectification that takes place.

It is difficult to draw up any statement as to the results so far arrived at in the measurement of rectifier performance for the reasons already indicated above, namely, that the bulk of the measurements refer to special cases and are not recorded in a form suitable for generalisation. The following qualitative conclusions appear to be well-established :—

(*a*) The measured performance is consistent with data derived from the static characteristics of the valve concerned, at least up to a million cycles in frequency. [DAVID claims to have extended this conclusion to about twenty million cycles, but additional confirmation of this would be desirable in view of the fact that a paper by W. E. BENHAM (see Abs. 4.3/19) throws some doubt on the matter.]

(*b*) An exponential representation of the grid-current characteristic can be used for the analysis of the grid-circuit rectification of signals up to an amplitude of about one volt.

(*c*) A parabolic representation of the anode-current characteristic can be used for the analysis of anode circuit rectification of signals up to an amplitude of 2 or 3 volts.

(*d*) In general, grid-circuit rectification will give a greater output than anode circuit rectification, but much depends on the actual circuit conditions.

(*e*) The grid-circuit rectification of telephony modulated continuous waves may, and in general will, involve some degree of modulation frequency distortion. This is also true of anode-circuit rectification in some cases (e.g., with high resistance loads) but in general the degree of distortion can be made less than in grid-circuit rectification.

(*f*) Grid-circuit rectification is not effective for the reception of heterodyne or telephony modulated waves at wavelengths longer than two or three thousand metres.

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Where grid-circuit rectification is apparently being used for long wave heterodyne or supersonic-heterodyne reception, it is probable that anode-circuit rectification conditions are set up by the agency of the local oscillator.

It will be noticed that no mention is made of crystal rectification. This subject is considered to be of little practical interest, since, in spite of the apparent sensitivity of crystal detectors where a fixed input voltage is concerned, their energy transformation efficiency compares very unfavourably with that obtainable by the proper use of valves. Moreover, their inherent variability makes them unsuitable for quantitative measurement.

BIBLIOGRAPHY

1. **A Study of Heterodyne Amplification by the Electron Relay :** E. H. ARMSTRONG : *Proc. Inst. Radio Eng.*, 1917, **5**, 145-59.

This paper describes an early but very important investigation of heterodyne reception. The input and output quantities were measured by means of crystal rectifiers, radio-frequency components being separated out in the output circuit by using an audio-frequency transformer with a suitable arrangement of by-pass condensers. The chief result obtained is that there appears to be a fairly pronounced optimum value for the local oscillator intensity.

This appears to be due to what the author calls " over-loading " the valve, i.e., excessive lowering of the grid potential by the local oscillator, so that anode-current characteristic curvature comes into effect. It is further shown that the ratio of the output for optimum heterodyne to that for equal heterodyne increases as some inverse power of the signal E.M.F.

The optimum condition realised by Armstrong in this work was apparently due to the fact that grid-circuit rectification was used, causing a lowering of mean grid potential which, for large local oscillator amplitudes, brings an opposing anode-circuit rectification into play. This optimum must not be confused with that due to anode-current saturation (see Abs. 2.3.2/14). The author does not explain the effect mentioned at the end of the preceding paragraph, but it follows from two results arrived at in this experimental work, namely, that the output with equal heterodyne is proportional to the square of the signal, and with optimum heterodyne to the first power of the signal.

It must be remembered that a very early type of valve was used in this work.

(See also Abs. 1.2.4/1, 2.3.2/2.)

2. **Detection Coefficient of Thermionic Amplifiers :** H. J. VAN DER BIJL : *Phys. Rev.*, 1919, **13**, 311-14.

The chief interest in this paper lies in the method of measurement used. The output telephone audio-frequency current derived from the valve rectification of a modulated continuous wave was measured by audible comparison with that derived from an audio-frequency generator, the large and measurable output from the latter being attenuated in a known manner to equality with the sound to be measured.

This would not be as accurate as a bridge method, but would have the advantage of comparative simplicity.

(See also Abs. 2.3.2/4.)

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3. Efficiency of a Single Thermionic Valve as a Detector : E. O. HULBURT and G. BREIT : *Jahrb. drahtl. Telegr.*, 1921, **17**, 119-32.

The paper is confined to the case of anode rectification of a pure continuous wave, and is devoted to the experimental checking of the analytically determined value of a so-called " detection coefficient " (which is defined as the limiting value of anode current change over the square of the input voltage, when the latter tends to zero).

A condenser potential divider was used for the determination of input voltages, and the output current change was measured by means of an electrometer connected across a non-inductive (liquid) resistance. A tuned circuit was included in the anode circuit, apparently for the purpose of checking the theoretical formulæ appropriate to such a case.

The method of measurement used would probably not be applicable at very high radio-frequencies.

(See also Abs. 2.3.2/7.)

4. The Thermionic Triode as Rectifier : E. B. MOULLIN and L. B. TURNER : *J. Instn. Elect. Eng.*, 1922, **60**, 706-19.

Grid and anode-circuit rectification of continuous and heterodyned continuous waves is considered, both theoretically and experimentally. For the latter purpose, 90 cycle alternating current is used, the authors making the assumption that there is no frequency effect in the valve itself.

Anode-circuit rectification is analysed on the basis of a parabolic form for the characteristic, and is found experimentally to conform very closely to this. Grid-circuit rectification is analysed on the same basis, which is only valid for rather small input voltages. The effect of varying the " grid-leak " and the grid coupling capacity are shown experimentally.

Some mention is made of the time-constant effect in the grid circuit.

The now well understood fact that with very large input voltages the lowering of the mean grid potential by the signal may give rise to an opposed anode-circuit rectification effect is demonstrated by an experimental curve.

The paper gives a very systematic treatment of the subject and contains much useful data, though these might not be quantitatively applicable to present day valves.

(See also Abs. 2.3.2/8.)

5. The Dynatron : A. W. HULL, E. F. HENNELLY, F. R. ELDER : *Proc. Inst. Radio Eng.*, 1922, **10**, 320-43.

The dynatron is a valve having two concentric grids and an anode. The grid nearer the filament functions as a control grid in the ordinary way. The anode is used merely as a source of secondary emission or " impact electrons," which are collected by the second grid (called the " dynode"). The valve is thus a secondary emission negative resistance device. The negative resistance characteristic is utilised for the generation of local oscillations for heterodyne reception.

From the point of view of the present section, the main interest of this paper lies in the experimental method used to give a qualitative confirmation of the analysis of the behaviour of the valve. Telephonic frequencies were used in the place of the normal radio-frequencies and oscillographic records taken of the current variations in the constituent circuits.

No discussion is given as to the possible effect of inter-electrode capacities in modifying the radio-frequency performance of the apparatus, as compared with the demonstrated telephonic frequency performance.

(See also Abs. 2.3.3/1.)

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6. Detecting Efficiency of the Resistance-Capacity Coupled Amplifier to 6,000 metres : W. G. BROMBACHER : *Phys. Rev.*, 1922, **20**, 433-40.

The results obtained in this paper show that data obtained from static valve characteristics can be used to predict the performance at radio frequencies. The formula considered refers to a particular two-stage resistance amplifier. Continuous wave rectification only is considered, the input voltages being obtained by means of a condenser potential divider.

The method employed for the measurement of the input voltage appears to be practicable at the low radio-frequencies considered, but would probably not be so at higher radio-frequencies.

(See also Abs. 2.3.2./9.)

7. An Experimental Study of Processes of Detection of High Frequency Oscillations : R. DUBOIS : *Onde Élect.*, 1924, **3**, 285-98, 347-63.

This paper describes an oscillographic study of the rectification of pure continuous waves and continuous waves modulated with a square wave form by means of a " tikker." Attention is mainly devoted to the distortion of the modulation wave form. The experimental results show that grid-circuit rectification may involve a diminution in the amplitudes of higher harmonics in the modulation wave form. The oscillographic records give qualitative confirmation of the effect predicted, and show very clearly that anode-circuit and crystal rectification are free from it. In conclusion, the author emphasises the fact that there appears to be no important difference between the dynamic and static characteristics of valves.

The experimental results are not presented in a form which facilitates general application to design, but they are consistent with the conclusions arrived at by DAVID, COLEBROOK and others. It appears, however, that the author has not realised that the resistance involved in the time constant effect is not that of the grid leak alone, but that of the grid leak in parallel with the grid-filament resistance.

(See also Abs. 2.3.2/12.)

8. The Rectifying Detector : F. M. COLEBROOK : *Experimental Wireless*, 1925, **2**, 330-4, 394-403, 459-68.

This paper contains numerous measurements of the continuous-current output, with various pure resistance loads, given by crystal rectification of pure continuous waves. The input voltages were obtained by using the fall of potential in the heater element of a non-contact thermo-junction. Measured values of effective rectified voltage and rectifier internal resistance are deduced from these results. It is shown that the effective internal resistance of, say, a galena detector will depend greatly on the amplitude of the applied signal voltage, varying from a hundred ohms or less to several thousand ohms as the amplitude is reduced to zero. In general, the rectified voltage will be an appreciable fraction of the amplitude of the applied alternating voltage. Measurements are also given of the effective output continuous rectified voltage with an infinite output-circuit resistance (potential rectification). This may be as much as 80 to 90 per cent. of the applied alternating amplitude (peak value). A valve with a negative mean grid potential was used for these latter measurements.

(See also Abs. 2.1/3.)

9. The Valve Reception of Modulated Waves, with Retro-action : C. KUHLMANN : *Jahrb. drahtl. Telegr.*, 1925, **25**, 95-109.

This paper is concerned with the combined effect of rectification and input circuit damping, the latter being controlled by retro-action, on the output derived from a modulated radio-frequency inducing field. The output

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measurements were made by means of a tuned vibration galvanometer of the Einthoven type, as in the earlier paper by the same author, and were confined to a single radio frequency, modulation frequency combination (6,000 metres, 7,500 cycles/sec.). The author states that, with such a long carrier wave length, reduction of the input circuit damping will not in itself lead to an increased modulation frequency output, since the latter is limited by the side band amplitude which, owing to the selectivity of the reduced damping effect, will not be much increased thereby. The curves, however, show an increased modulation-frequency output with reduced input circuit damping, but this may be attributed entirely to the establishment of more favourable rectification conditions owing to the increased amplitude of the carrier wave.

The treatment and the conclusions are not very clear, owing to the fact that too many variable elements are involved in the process. The main conclusion, however, that retro-action is less effective with long-wave than with short-wave telephony seems to be consistent with general experience. This matter requires further investigation.

(See also Abs. 2.3.3/10.)

10. **The Valve Reception of Modulated Waves without Retro-action :** C. KUHLMANN : *Jahrb. drahtl. Telegr.*, 1925, **25**, 70-82.

This paper is concerned with grid-circuit rectification of modulated waves (wave-length 6,000 metres, various modulation frequencies). A theoretical analysis is given, assuming an exponential form for the characteristic. The measurements of performance were made with a resonant vibration galvanometer of the Einthoven type, a very useful means of selecting the particular audio-frequency component under observation. They record the variation of the output, for constant amplitude input, with the magnitude of the " grid-leak " and the grid condenser, such measurements being repeated at a large number of modulation frequencies. The frequency distortion effect due to the variation of the audio-frequency impedance of the grid leak-condenser combination is very clearly demonstrated. It is shown that too small a coupling capacity reduces the output owing to a fall of radio-frequency potential across it, while too large a capacity reduces the output owing to the shunt effect on the grid-leak.

It should be pointed out (and this criticism applies equally to Colebrook's theoretical demonstration of the same effect) that the case of a maintained input potential alone is considered. Very considerable modifications of the optimum grid condenser conditions will be found when, as is so often the case, the input is a resonant voltage across a tuned circuit.

(See also Abs. 2.3.3/9.)

11. **Detecting Characteristics of Electron Tubes :** H. M. FREEMAN : *Proc. Inst. Radio Eng.*, 1925, **13**, 611-23.

The author contemplates studying the action of valve rectifiers with a view to the improvement of the design of valves to be used for this purpose. The present paper describes measurements made to check the simple approximate formula for grid rectification for comparatively small input voltages (parabolic characteristic). A fixed modulated radio-frequency input voltage was used (the paper does not say how it was measured) and the modulation-frequency output was measured by a simple slide-back valve voltmeter. The mean grid voltage was the principal variable considered.

The curves obtained are only of interest as showing that the basic approximate formula can be checked experimentally.

(See also Abs. 2.3.2/16.)

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12. **The Rectification of Small Radio-frequency Potential Differences by means of Triode Valves :** F. M. COLEBROOK : *Experimental Wireless*, 1925, **2**, 865-77, 946-57 ; 1926, **3**, 34-40, 90-5.

This paper deals very comprehensively with the various types of valve rectification. Grid-circuit rectification is analysed on the basis of an exponential representation of the grid current-grid voltage characteristic and anode-circuit rectification on the basis of a parabolic characteristic.

Measurements are given in confirmation of the analysis for the rectification of pure continuous waves, the input voltage being obtained by using the fall of potential in the heater of a non-contact thermo-junction ammeter. It is shown that the assumption of an exponential grid-current characteristic is valid to a close approximation for ordinary small receiving valves, for a range of grid voltages from – 1 to 0, which covers a useful working range. It is also shown that with suitable components (grid-leak and grid condenser) the process appears to be independent of the frequency of the applied voltage. (This applies only to pure continuous wave rectification.)

Analysis and confirmatory measurements are also given for the determination of the appropriate value of the grid condenser. It must be emphasised, however, that this analysis and the measurements refer to the assumption of a given input potential.

A comparison, based on measurement, is made between grid- and anode-circuit rectification for pure continuous waves. It is shown that, from the point of view of maximum output power obtainable from a given alternating input potential difference, the former has from 50 to 100 times the sensitivity of the latter (a ratio of the order of 10 to 1 expressed in terms of output circuit continuous current).

This latter conclusion is limited to the particular type of rectification considered, and requires fuller investigation in relation to other applications. A very significant feature of this paper, to which attention is drawn, is that data obtained from the static characteristics of the valves considered are valid as a basis for calculating the behaviour of the valves at high radio-frequencies, indicating that there is no important difference between the dynamic and static characteristics of valves.

(See also Abs. 2.3.2/15.)

13. **Anode Rectification :** M. VON ARDENNE : *Jahrb. drahtl. Telegr.* 1927, **29**, 82-88.

Analysis and measurements are given for the anode-circuit rectification of pure continuous waves with resistance in the anode circuit. The measurements (for the most part) were made at an audio-frequency (800 cycles/sec.) input voltages being measured by a string electrometer. The output continuous current curves are drawn for constant anode-circuit resistance and input voltage, and only show the variation with mean grid potential. They also show that the output is greatly increased, as predicted by analysis, by effectively short circuiting the anode circuit resistance with a capacity.

The paper is of rather limited application and cannot be described as a systematic and thorough study of its subject.

(See also Abs. 2.3.2/25.)

14. **A Theoretical and Experimental Investigation of Detection for Small Signals :** E. L. CHAFFEE and G. H. BROWNING : *Proc. Inst. Radio Eng.*, 1927, **15**, 113-53.

The analytical side of this paper is abstracted in Section 2.3. The conclusions therein arrived at were checked experimentally, as far as modulated C.W. is concerned, by the rectification of a radio-frequency potential difference modulated to a known extent with a thousand cycle note. The method used

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is a bridge arrangement in which the rectifier output current is balanced against a known modulation-frequency current derived from the modulator source.

The actual results refer mainly to soft valves and to particular cases, and they only serve to check, approximately, the theoretical analysis (which is limited to the case of small signals, i.e., a parabolic characteristic is assumed) for the particular case considered. The analytical presentation of the subject employs the conceptions of effective rectifier output voltage and internal resistance, but the paper does not give any systematic application of these ideas to design.

(See also Abs. 2.1/4, 2.3.2/24.)

15. **Voltage Detection Coefficient :** E. L. CHAFFEE : *Proc. Inst. Radio Eng.*, 1927, **15**, 946-57.

This is a continuation of the former paper by the same author. Here the conception of effective rectifier output voltage (in the particular case of the rectification of a modulated continuous-wave potential difference) is made more explicit. It is embodied in a quantity which the author calls the " voltage detection co-efficient." This quantity is defined in a manner which implies that it is proportional to modulation percentage. It is further assumed that this effective output voltage will be independent of the modulation frequency load. The idea of effective rectifier internal resistance is also used. Curves are given showing the variation of effective rectified voltage with mean grid potential for constant input voltage amplitude, modulation percentage and modulation frequency. The measurements were made by the bridge method referred to in the earlier paper.

The two assumptions referred to above will not always be accurately valid. The effective modulation-frequency voltage may vary appreciably with load. In dealing with grid-circuit rectification, the subject is unnecessarily complicated by introducing the characteristics of a resonant circuit input. It is preferable to treat the rectifier as a four terminal network having a certain input impedance, the effect of which on any resonant input circuit can be considered separately. The paper is nevertheless to be commended as an attempt to systematise the subject, and the method of measurement employed has useful possibilities.

(See also Abs. 2.3.2/26.)

16. **Grid Rectification :** Y. B. F. GROENEVELD, B. VAN DER POL JR., and K. POSTHUMUS : *Jahrb. drahtl. Telegr.*, 1927, **29**, 139-47.

This is an important contribution to the theory of grid rectification. An exponential form of representation for the grid-current characteristic is used, and is related to the physics and to the geometry of the valve. Details of the measurement method are not given, but it is shown that the rectification of pure continuous-wave signals can be satisfactorily analysed over an important range of voltage variation in terms of an exponential characteristic. It is shown that the effective grid-voltage change due to rectification approaches the peak value of the signal as an upper limit for large signal voltages.

(See also Abs. 2.3.2/20.)

17. **Detection by Grid Rectification with the High Vacuum Triode :** S. BALLANTINE : *Proc. Inst. Inst. Radio Eng.*, 1928, **16**, 593-613.

This paper is more fully considered in Section 2.3. Its interest in the present Section is mainly the method suggested for determining the audio-frequency distortion associated with the grid rectification of telephony modulated signals. It is shown that the magnitude of this effect can be demonstrated by applying audio-frequency voltages to the input terminals

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of the rectifier and measuring the audio-frequency output. (The output measuring arrangement is not described.) Experimental curves are given, showing that a frequency distortion of the order of 50 per cent. to 60 per cent. may arise, even in comparatively short wave reception, owing to this cause.

(See also Abs. 2.3.2/29.)

18. **Quality and the Anode Rectifier :** A. L. M. SOWERBY : *Wireless Wld.*, 1928, **22**, 87-90.

Results are given of the measurement of the rectified potential difference across a resistance in the anode circuit of a valve having measured (50 cycle) alternating voltages applied to the grid. The curves show a marked superiority in high voltage factor valves with high anode voltages in respect of the quantities actually measured. On the basis of these measured curves, various deductions are made with respect to the behaviour of such rectifying systems in relation to broadcast telephony.

It is doubtful whether the curves can legitimately be used for this purpose, owing to the differences in the operating conditions contemplated. Nevertheless, the curves have some interest in themselves, apart from the deductions drawn from them.

19. **A Study of the Rectification Efficiency of Thermionic Valves at Moderately High Frequencies :** W. E. BENHAM : *Phil. Mag.*, 1928, **5**, 323-4.

Measurements of the change of anode current in a diode with applied measured radio-frequency voltages over a wave-length range from about 70 to 17 metres show a marked variation of a systematic character in rectification efficiency, which variation cannot apparently be attributed to anything except the valve.

This conclusion is inconsistent with that arrived at by DAVID, who worked down to 14 metres. There is some possibility of doubt as to the behaviour of a Dolezalek electrometer at these frequencies (this being the input measuring instrument employed by BENHAM). The inductance of the fine suspension filament, for instance, may introduce error. Further, the question of wave form enters into this matter and no special precautions seem to have been taken in this respect. The matter would appear to require further investigation.

(See also Abs. 2.3.1/5.)

20. **Valve Detection :** P. DAVID : *Onde Élect.*, 1928, **7**, 313-58.

This paper contains three features of special interest in relation to the measurement of the performance of rectifiers.

(*a*) Measurements of continuous current output with grid-circuit rectification are made down to 14 metres wave length, using a Wilson gold leaf electroscope as the input measuring instrument. Within the limits of experimental error, the rectifier performance is found to be substantially independent of the frequency down to this limit.

(*b*) Measurements of beat frequency output are made, with grid and anode-circuit rectification. The author enlarges on the difficulty of this measurement in view of the necessity for eliminating continuous and radio-frequency components from the total output potential variation. The separation is effected by transferring the output potential to be measured to the measuring amplifier by a resistance-capacity connection of the usual kind, the resistance being shunted by an acceptor circuit tuned to the radio-frequency.

With the above apparatus, the author is able to demonstrate the falling off of grid-circuit rectification efficiency with increasing beat or modulation frequency.

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(*c*) The author points out that since grid-circuit rectification is inefficient for very high audio and ultra-audio modulation frequencies, it should be quite unsuited to supersonic-heterodyne reception. Nevertheless, the grid-circuit rectification connections are commonly used for this purpose, and appear to be effective. He resolves this apparent discrepancy by showing that in actual fact, anode-circuit rectification will take place in this case, the mean grid potential being automatically lowered to an appropriate value by grid-circuit rectification of the local oscillator E.M.F. This conclusion is verified by measurement, using the apparatus described above.

It may be pointed out in criticism of this method that the audio-frequency output is also of more or less complex wave form, and this fact is not taken into account. If, as appears to be the case, the measuring amplifier is calibrated with a sine wave audio-frequency source, the measurements will contain a wave form error.

(See also Abs. 2.1/5, 2.3.2/21.)

21. **Performance of Valve Detectors :** W. B. MEDLAM and V. A. OSCHWALD : *J. Inst. Wireless Tech.*, 1929, **1**, 147-223.

The rectification performance of triode valves with resistance loads in the anode circuits is examined. For pure continuous wave measurements, a sharply tuned circuit was loosely coupled to an oscillator, and the voltage obtained across the tuning condenser utilised as the input voltage. A thermionic voltmeter connected directly across the condenser was used to measure the voltage. Modulated input voltages were obtained by loosely coupling the tuned circuit to two radio-frequency oscillators, the modulation frequency being the difference between the two radio frequencies. The common coupling between the two oscillators necessarily involved in this arrangement was neutralised by a similar independent coupling of opposite sense, in order to avoid the locking of the two oscillators at the lower beat frequencies. The output modulation frequency potential differences were measured by a thermionic voltmeter. Of the harmonics in the output circuit, it was found that only the double modulation frequency was appreciable compared with the fundamental. This was eliminated by a tuned " rejector " circuit, enabling measurements of the fundamental only to be made.

The distinctive experimental features of this investigation are the filtering of the audio-frequency output potentials and the use of the heterodyne arrangement as a source of modulated radio-frequency potential differences. The analyses given in support of the statement that the input voltages so obtained do not differ in any essential respect from a pure tone modulated wave is not entirely convincing, but the method is evidently a useful one for investigating the behaviour of rectifiers or of receivers in general.

(See also Abs. 2.3.1/10.)

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4.4. AUDIO-FREQUENCY AMPLIFIERS

4.4.1. Voltage Amplification

The essential conditions which need to be fulfilled in order that a satisfactory measurement of the amplification given by an audio-frequency amplifier may be obtained are not dissimilar from those which exist in the radio-frequency case. These conditions have already been outlined in Section 4.2, and it is only necessary therefore to examine methods which have been adopted for this purpose which are peculiar to the audio-frequency case. The limitations imposed upon the measuring apparatus are :—

(i) That the source of the input potential shall be such that it does not modify the normal behaviour of the amplifier.
(ii) That the nature of the measuring apparatus which records the output shall be such that it does not modify the normal behaviour of the amplifier.
(iii) That no direct retro-action effect shall take place between the input and output circuits due to the presence of the measuring apparatus adopted.

Five methods of measurement have been used, namely :—

(i) Direct methods.
(ii) Substitution methods.
(iii) Null methods.
(iv) Bridge methods.
(v) Oscillographic methods.

In a direct measurement, the input voltage is applied by a mutual inductance or a potentiometer, and the output is measured by a thermo-junction ammeter, an electrometer, a vibration galvanometer, or a valve voltmeter. Although the determination of mutual inductance at audio-frequencies is comparatively easy, this method has not been used much owing to the comparatively large reactance introduced across the input terminals by the inductance of the secondary circuit of the mutual. The potential difference across this secondary-circuit reactance will be materially modified by the input impedance of the amplifier. If the source of input potential is derived from the potential drop across a known resistance carrying a known current, the value of this resistance must be low (less than 100 ohms) if retro-action effects are to be negligible.

It is preferable to express the amplification factor as the ratio of the voltage delivered to the power valve to the applied input voltage. In this case the voltage delivered to the power valve

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may either be calculated from the current output and the known constants of the valve, or preferably the current-indicating instrument in the anode circuit of the power valve may be calibrated in terms of the grid voltage, by means of a potentiometric input. It is unwise to connect a voltage recording instrument directly across the grid and filament of a power valve, since the loading effect of this instrument may materially modify the behaviour of the amplifier and, even if a recording instrument is inserted in the anode circuit of the power valve, care must be taken that its load does not materially modify the effective input impedance of the stage. This means that a large resistance or inductance cannot be inserted in the anode circuit. A thermo-junction ammeter or a vibration galvanometer of low resistance is permissible, or an electrometer, or a valve voltmeter connected across a resistance of not more than about 500 ohms. These conditions do not appear to have been fully appreciated in all measurements that have been made; for instance, M. VON ARDENNE (see Abs. 4.4.1/10) inserts a resistance of 6,000 ohms in the anode circuit and measures the voltage across it by means of an electrometer. The resistivity of a thermo-ammeter is far less than that obtained with a vibration galvanometer but the sensitivity of this instrument depends upon frequency. The most satisfactory all round instrument is the valve voltmeter, since the sensitivity is reasonably good and the calibration is independent of frequency. In any direct measurement, care must be taken to screen the input oscillator, the potentiometer or attenuation network, the amplifier and the output recording mechanism. In many cases sufficient care has not been taken to ensure that the effects of retro-action are reduced to a negligible amount.

A substitution method is one in which a definite but unknown output is obtained, after which the same output is obtained by the direct connection of the output device to the input oscillator. The SMITH-NAPIER (see Abs. 4.4.1/1) and PIRANI (see Abs. 4.4.1/2) methods may be classified under this heading; in both cases a telephone comparison is made between the output obtained in the normal manner and the direct connection to the telephone. Such methods usually possess an upper amplification limit, and do not take into account distortion effects. The accuracy of any such telephone balance is limited by physiological considerations and at the best is only about 5 per cent.

A null method is one in which zero output is obtained by applying an E.M.F. to the telephone, or grid of the last stage, of equal magnitude to the actual E.M.F. and having a phase difference of 180°; if the value of this E.M.F. is known, the output voltage is determined. An example of this type is the method due to WILLANS (see Abs. 4.4.2/7). The method is open to criticism on the grounds that the normal output conditions may be seriously modified by the presence of the auxiliary apparatus required to give the null condition. If a telephone balance is adopted, balance will be

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obtained for the fundamental frequency only, the harmonics due to distortion remaining unbalanced. This is an advantage not possessed by methods in which the R.M.S. value of the output is obtained. Although there is a divergence of opinion regarding the accuracy with which a telephone balance of the fundamental frequency can be obtained in the presence of harmonics, there seems little doubt that methods which neglect the harmonic constituents of the output are preferable to R.M.S. determinations. This advantage, however, is not sufficient to justify the general use of this method, owing to the altered condition of the amplifier due to the presence of the auxiliary measuring apparatus.

The bridge method possesses similar advantages to the null method, in so far as the balance is obtained for the fundamental frequency only, but it has the added advantage of not reducing the alternating component of the anode current of the last stage to zero. This method is satisfactory for reasonable amplifications, but it is clear that in the measurement of large amplifications elaborate precautions must be taken to ensure freedom from retroactive effects introduced by the bridge connections. Although a little information on this subject has been published by DIAMOND and WEBB (see Abs. 4.4.1/11), detailed information on the use of the bridge method has not been published, and it is desirable that the limitations of this method be examined in order that it may be compared with direct methods.

The oscillographic method is useful if a rough determination of amplification only is desired. If the output is applied to one pair of oscillograph plates, while the other pair are connected across a resistance which is a known multiple of the input-potentiometer resistance, the ratio of the axes of the Lisajous figure so produced gives the amplification factor, while the elliptical form of the figure gives the phase. To obtain the complete amplification-frequency characteristic of an amplifier, DIAMOND and WEBB developed an ingenious method in which the input frequency is continuously varied by rotating the tuning condenser of the oscillator; by mechanically connecting this condenser to a voltage "sweep" device, the cathode ray spot was made to move horizontally in such a manner that one end of its movement corresponds to a low frequency, while the other end corresponds to a high audio-frequency. By this means a fluorescent area is obtained, since the other pair of plates is connected to the output, and the bounding edge of this area gives the required amplification-frequency characteristic. It is necessary to apply a constant input E.M.F. at all the audio-frequencies, which is done by using a heterodyne oscillator.

4.4,2. TRANSFORMER PROPERTIES

A bridge method is usually employed to determine the constants of the inter-valve transformer. Usually the measurement consists in observing the change in effective inductance and resistance of

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a non-inductive resistance when shunted by the primary of the transformer (see Abs. 4.4.2/2). The value of the primary inductance and resistance is dependent on the magnetising current (see Abs. 4.4.2/11) and also on the amplitude of the alternating current, and consequently it is important that these currents are made similar to those occurring in practice. Neglect of these points has rendered unreliable many results obtained by the use of the bridge method. Shielding devices should be used for the resistances and the leads of the bridge, particularly where the currents in the arms are large. Other methods of measurement, such as volt-ammeter, three-voltmeter (see Abs. 4.4.2/11), and substitution, have been used and in many cases such measurements give the information required more readily and with less apparatus than the bridge method, but the accuracy of such measurements is usually not high.

An enormous quantity of information is available regarding the measured performance of audio-frequency amplifiers, but a review of the whole of it reveals enormous overlapping of work, repetition, and a misdirection of effort. Little can be gained by obtaining innumerable characteristic curves of amplifiers unless, as a result of such curves, general design conclusions are obtained.

The whole of the work on audio-frequency amplifier design has been directed towards obtaining a linear-frequency characteristic. It should be pointed out, however, that this will only result in a linear overall performance characteristic if used in conjunction with a loud speaker which is itself effectively linear in response. Since loud speakers do not at present possess such characteristics, three alternatives suggest themselves :—

(i) Improvement in the design of loud speakers, directed to obtain a uniform frequency-response. This is the ideal solution, and progress is being made in this direction.

(ii) Correction of the loud speaker response by suitable electrical filter circuits, connected between the loud speaker and the amplifier.

In the above two cases, an amplifier having a uniform frequency characteristic is required.

(iii) Design of the amplifier to compensate for the non-uniform response of the loud speaker. In this case, the amplifier and loud speaker would have to be regarded as an inseparable combination. This alternative cannot be recommended for that reason.

BIBLIOGRAPHY

4.4.1. Voltage Amplification

1. **On the Measurement of Amplification given by Triode Valves at Audible and at Radio-frequencies :** F. E. Smith and H. G. Napier : *Proc. Phys. Soc. Lond.*, 1920, **32,** 116-132.

A valve oscillator supplies a primary coil which is coupled to two secondaries rigidly fixed together at right angles, but capable of rotation with

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respect to the primary. When the mutual inductance is a maximum for one coil it is a minimum for the other, and by calibrating this mutual inductance for various positions, the ratio of the E.M.F.s produced in the secondaries is known. The two coils are connected alternately to the input of the amplifier and to the grid of the last stage by means of a six-contact switch, the position of the secondaries being adjusted until the signal strength heard in the telephones is the same for both positions of the switch. The ratio of the mutual inductances with respect to the primary then gives the voltage amplification. The pointer connected to the rotating secondary coils may move over a scale which is directly calibrated in amplification ratio.

The method is very quick and useful for the measurement of the amplification of one stage at audio-frequencies, but the accuracy of calibration of the mutual inductance for voltage ratios above about 30 becomes very inaccurate. Although the method is useful for a quick direct measurement of one stage, it is of no value as a general method owing to its limited range.

SMITH-ROSE gives a detailed description of the method (see Abs. 4.4.2./4.), together with experimental results obtained with various transformer coupled stages.

(See also Abs. 4.2/1.).

2. **Amplification Ratio of Radio-frequency and Audio-frequency Amplifiers :** M. PIRANI : *Jahrb. drahtl. Telegr.*, 1920, **16**, 2-30.

An audio-frequency source feeds a series circuit consisting of two resistances. The amplifier input is tapped across one of them and the output is given on a telephone. The telephone is now disconnected from the amplifier and is tapped across the second resistance, which is adjusted until a similar signal is obtained. The ratio of the resistances then gives the ratio of the volts across the telephone to the input voltage.

The limiting accuracy with which any telephone comparison can be obtained is of the order of 5 per cent., since the human ear is very insensitive to small changes in intensity. The method is essentially a substitution measurement in which a definite output is obtained in two ways.

(See also Abs. 4.2./2.)

3. **On the Voltage Amplification Ratio of the Triode Valve Resistance Amplifier :** H. NUKIYAMA and T. KUWASHINA : *Tôhuku University Technical Reports*, 1924, **4**, 1-17.

An alternating current potentiometer, supplied by an audio-frequency valve oscillator, is used for the measurement of the amplification ratio of a single stage with a resistive load in the anode circuit. The dependence of the amplification upon the value of this resistance is examined theoretically, and experimental results are given in confirmation of the theory. The loading effect of a grid-circuit rectifier is examined, and it is shown experimentally that the amplification falls as the output voltage feeding the rectifier increases : this is due to the damping produced by the grid current. A short bibliography is given.

4. **Some Thermionic Valve Problems :** E. V. APPLETON : *Rep. Brit. Ass.*, 1925 ; *Electrician*, 1925, **95**, 266-7.

(See Abs. 1.2.1/17.)

5. **Distortion in Amplifiers :** G. W. SUTTON : *Wireless Wld.*, 1925, **17**, 156-9.

The results of measurements on resistance-capacity-coupled and transformer coupled stages are given in graphical form ; the method of measurement,

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employing a simple alternating current potentiometer, is similar to that adopted by Pirani. The defective amplification-frequency characteristic obtainable by using too large a step-up transformer ratio is clearly shown, and the necessity for a large primary inductance is pointed out.

(See also Abs. 3.1.2/3.)

6. **The Best Working Condition for a Resistance Amplifier :** M. MERCIER : *Onde Élect.*, 1926, **5**, 413-24.

An account is given of an investigation dealing with the various factors affecting the amplification of a two-stage resistance-capacity-coupled amplifier. The effects of varying the anode voltage, anode-circuit resistance, and the filament current are shown graphically. The tests were made by changing the steady potential of the first grid and examining the change in current in the anode circuit of the second stage. The grid potential of the second stage was adjusted by batteries, no grid leak or condenser being used.

Such a method of test can only give the operating performance at very low audio-frequencies, and is of no value whatever as a means of examining the properties of an amplifier, since all capacity effects are neglected in the measurement.

7. **The Performance of Amplifiers :** H. A. THOMAS : *J. Instn. Elect. Eng.*, 1926, **64**, 253-68, 268-78 ; *Experimental Wireless*, 1926, **3**, 15-18 ; *Elect. Rev. Lond.*, 1926, **98**, 73-4.

Section I of this paper describes the application of the National Physical Laboratory method of measuring amplification to audio-frequency measurement. The method consists in applying a known small E.M.F. to the input of the amplifier by means of a non-inductive resistance potentiometer. The output is measured by a vibration galvanometer, which is calibrated by means of another potentiometer. A wide range of amplification is thus measurable with constant accuracy. Careful screening methods are adopted to avoid undesired induction effects.

(See also Abs. 1.2.7/6, 1.4.1/16, 3.4/9, 4.2/8, 4.5/7, 4.6/1.)

8. **Modern Amplifier Performance :** P. K. TURNER : *Wireless Wld.*, 1926, **19**, 861-2.

A comparison of the performance of three different types of audio-frequency amplifiers is given, amplification-frequency characteristics being shown graphically. The method of measurement is not given.

The most uniform amplification-frequency characteristic was given by a resistance-capacity-coupled amplifier consisting of two valves having voltage factors of 20 each, in conjunction with anode resistances of 1 megohm, coupling condensers of 0·005 micro-farad, and grid leaks of 5 megohms each. It is stated that the effective frequency range is from 20 to 10,000 cycles per second, the maximum amplification being 620.

This paper is purely descriptive of the particular cases considered, and no attempt at generalisation is made.

9. **Notes on the Testing of Audio-frequency Amplifiers :** E. DICKEY : *Proc. Inst. Radio Eng.*, 1927, **15**, 687-706.

The paper deals chiefly with a method of measuring one audio-frequency stage. The input voltage is supplied by a potentiometer which is supplied with current from an audio-frequency source, the current being measured by a thermo-junction ammeter. The bulk of the paper is devoted to a careful description of a valve voltmeter used to measure the output. Lastly, a description is given of a special curve-drawing apparatus in which the amplification-frequency characteristic is plotted directly by means of a

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rotating drum geared to the audio-frequency oscillator condenser and thus rotating as the frequency is changed.

10. **Measurement of Voltage Amplification of Low-frequency Amplifiers :** M. VON ARDENNE : *Electrotech. u. Maschinenb.*, 1927, **45**, 25-8.

A method is described for the measurement of the amplification of one stage of audio-frequency amplification. The input is applied by means of a tapped potentiometer having a total resistance of 6,000 ohms, and the output is measured as a potential across a non-inductive resistance of 6,000 ohms inserted in the anode circuit of the stage amplification. An electrometer (see **A Single Thread Electrometer :** T. WULF : *Phys. Z.*, 1914, **15**, 250-4) is used as the measuring instrument for both the input and the output voltages. By using a very fine thread of diameter 2μ, and a small plate distance, effective potentials of the order of 0·5 volt could be measured with sufficient accuracy. The electrometer is first connected across the potentiometer input, and then across the resistance in the output circuit.

Examples are given of the application of the method to various types of apparatus, and the results obtained are shown to be in fair agreement with the simple theory of the resistance amplifier.

11. **The Testing of Audio-frequency Transformer-coupled Amplifiers :** H. DIAMOND and J. WEBB : *Proc. Inst. Radio Eng.*, 1927, **15**, 767-91.

Four methods of measuring amplification are given. The first two methods are essentially bridge circuits, using a Wagner earth, in which the amplifier output E.M.F. is balanced against an auxiliary voltage, the magnitude and phase of which is determinable with respect to the input voltage. An auxiliary valve with telephones in the anode circuit is used to effect this balance. In the third method, the auxiliary balancing valve is omitted, the last stage of the amplifier being used in its place. The fourth method is oscillographic. A heterodyne audio-frequency oscillator is used as the source ; such an oscillator gives a variable frequency with constant amplitude. This signal is applied to the input circuit of the amplifier, the output voltage being connected across one pair of deflecting plates of a cathode ray oscillograph. The other pair of deflecting plates is connected to a " sweep " voltage arrangement whereby the electron beam is deflected horizontally, the " sweep " voltage being so adjusted that the spot is at the extreme left when the signal frequency is a minimum, and at the extreme right when the signal frequency is a maximum. By this means, the pattern appearing on the screen gives the desired amplification-frequency characteristic. By adopting a slight modification of the " sweep " voltage device, this characteristic may be obtained to a logarithmic scale of frequencies.

This method does not give very accurate results owing to the limitation set by the dimensions of the image on the fluorescent screen of the oscillograph, but is very useful for rough quick measurements.

4.4.2. TRANSFORMER PROPERTIES

1. **Effect of Transformer Resonance :** K. MÜHLBRETT : *Arch. Elektrotech.*, 1920, **8**, 365-90.

(See Abs. 3.3/3.)

2. **The Performance and Properties of Telephonic Frequency Inter-valve Transformers :** D. W. DYE : *Experimental Wireless*, 1924, **1**, 691-9 ; and 1924, **2**, 12-21, 74-84.

The method of measurement described employs a bridge with four equal ratio arms of 1,000 ohms each. The measurements consist in observing the

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change in effective inductance and resistance of a non-inductive resistance of 500 ohms inserted in one arm, when shunted by the primary of the transformer. An Ayrton-Mather low-reading electrostatic voltmeter serves to read the secondary voltage of the transformer, this voltmeter being disconnected when the bridge observations are being made owing to the large effect produced on the effective primary inductance and resistance due to the capacity of this voltmeter.

(See also Abs. 3.3/5.)

3. **Telephone Transformers :** W. L. CASPER : *J. Amer. Inst. Elect. Eng.*, 1924, **43**, 197-209.

After a general discussion of the various uses to which transformers are put in telephony, and the requirements that these involve, the efficiency of a transformer is measured by comparison with an ideal transformer with no losses and the correct ratio in three cases : (1) When the circuits it couples are resistances ; (2) when one circuit is a resistance and the other a negative reactance ; (3) when one is a resistance and the other a positive reactance. In all these cases, the transformer is replaced by an equivalent network, and the variation of transformer losses with frequency is discussed.

4. **The Measurement of Low-frequency Amplification :** R. L. SMITH-ROSE : *Wireless Wld.*, 1924, **13**, 699-702, 731-74.

The second part of this paper gives results obtained with various single stages employing transformer coupling. The transformers used were of an early type and do not represent present day performance, but the characteristics given serve to show the important improvements in design that have been made.

It will be noticed that the amplification is dependent upon the method of connecting the windings ; the difference in amplification with various connections being due to the capacitative coupling which was so serious in all the early forms of transformer.

(See also Abs. 4.4.1/1.)

5. **The Impedance of Audio-frequency Amplifier Transformers :** S. CHIBA and S. KITTA : *J. Inst. Elect. Eng. Jap.*, 1925, No. 445, 733-8.

The impedance of audio-frequency inter-valve transformers is measured by the bridge method, a circle diagram being given representing the impedance at various frequencies. The magnitude of this impedance will depend upon the value of the magnetising component, the distributed capacities between the windings, and the current flowing in the secondary circuit of the transformer. The effect of this secondary circuit current is very important, since, even if the grid of the succeeding valve is negative, the secondary circuit will be loaded due to the input impedance of this valve. The manner in which the primary impedance depends upon the secondary circuit load is investigated.

6. **Audio-frequency Transformer Characteristics :** T. SPOONER : *Elect. J.*, 1926, **23**, 367-75.

(See Abs. 3.3/11.)

7. **Low-frequency Inter-valve Transformers :** P. WILLANS : *J. Instn. Elect. Eng.*, 1926, **64**, 1065-88, and *Experimental Wireless*, 1926, **3**, 435-439.

In this method, a bridge with Wagner earth is used in which one arm supplies the input voltage to the stage, and the other arm supplies an E.M.F. to a valve in parallel with the output ; the magnitude of this E.M.F. is the

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same as that of the output, while the phase difference is 180°. At balance, there is no current in the common output lead connected to the two anodes.

It is claimed that by this method the true amplification of the stage is obtained, but the method of measurement has been severely criticised, the grounds for such criticism being that the bridge circuit adopted does not represent the actual working conditions. It appears likely that a bridge measurement of this type may introduce circuit modifications which will give rise to effects not present in the normal operating condition.

(See also Abs. 3.3/14.)

8. **Amplifier Transformers :** L. MÜLLER : *Arch. Elektrotech.*, 1926, **16,** 219-50.

The second part of this paper deals with experimental work performed on 22 different transformers with a view to understanding their properties. The transformers selected for examination fall into three types, one type having a closed iron circuit, a second type having a straight core of iron wires and enclosed in a screen made of soft iron with a copper outside shield, and a third type having an open iron circuit.

An audio-frequency bridge arrangement with Wagner earth was used, and special screening devices were used for shielding the resistances and leads of the bridge.

The measurements refer to the transformer only and not to a stage of amplification using a transformer coupling.

(See also Abs. 3.3/15.)

9. **Resonance in Telephone Transformers :** J. GRANIER : *Rev. Gén. Elect.*, 1926, **20,** 789-94.

After a brief mathematical explanation, the author describes a simple bridge arrangement used to measure the transformation ratio of inter-valve transformers at audio-frequencies ; no special earthing devices are adopted to eliminate the effects of stray earth capacities.

The author notes that a transformer has two resonance points, one at a normal telephone frequency, and the other at a nearly inaudible frequency. This property had already been observed and analysed by DYE. The author states that this upper resonance point produces an excessive voltage amplification and leads to frequency distortion. In modern design, however, this property is now utilised to preserve the upper frequencies. Amplification frequency characteristics are given for the transformer, with various anode resistances in series with the primary winding. From these, it is shown that a sensibly uniform characteristic can be obtained over a large frequency range if a low impedance valve is used.

10. **On the Testing of Audio-frequency Transformers by means of the Cathode Ray Oscillograph :** M. KOBAYASHI : *J. Inst. Elect. Eng. Jap.*, 1928, No. **475,** 152-9.

Impedance-frequency and step-up-ratio-frequency characteristics are obtained by the use of a cathode ray oscillograph. The phase and magnitude of the secondary circuit output E.M.F. are obtained in terms of the primary circuit voltage, by analysing the elliptic figure on the oscillograph screen.

11. **The Measurement of Choke-coil Inductance :** C. A. WRIGHT and F. T. BOWDITCH : *Proc. Inst. Radio Eng.*, 1928, **16,** 373-84, Disc. 844-7.

A description is given of an investigation that emphasises the following facts :—

(i) The inductance of a choke coil depends upon the degree to which its core is magnetically saturated due to the direct current flowing through the winding of the choke coil.

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(ii) With a given direct current flowing through the winding of a choke coil, the inductance varies to a marked extent with the magnitude of the alternating current flowing through the winding. Methods of measurement which do not take into account the magnitude of the alternating current are, therefore, unreliable. It is in failing to take account of the magnitude of the alternating current that many bridge, ammeter-voltmeter, three-voltmeter, substitution, and fluxmeter methods of inductance measurement are unreliable.

(iii) The inductance for given conditions may be determined from the saturation curve of the iron. It is determined by the average slope of the saturation curve over the range within which the current varies.

Three modifications of the ammeter-voltmeter method of measuring inductance are presented :—

(1) A direct current of the required magnitude is passed through the choke, the current being measured by a thermo-junction ammeter, while the potential across the choke is measured by an electrostatic voltmeter. An alternating current is now super-imposed upon the direct current by means of a variable ratio transformer. The current is now equal to

$$\sqrt{I^2_{A.C.} + I^2_{D.C.}}$$

and the potential is equal to

$$\sqrt{E^2_{A.C.} + E^2_{D.C.}}$$

from which the inductance can be determined.

This circuit is applicable where only a few approximate measurements are to be made and where simplicity of connection is of the greatest importance.

(2) A disadvantage of the previous method is that, in order to obtain differences in current readings sufficient to permit of accuracy in the computed vector differences, comparatively large values of alternating current must be passed through the coils, probably larger than those met with in actual practice. This disadvantage may be avoided by connecting, in parallel with the thermo-junction ammeter, a circuit consisting of a large choke coil, batteries and a direct current ammeter. By suitably adjusting the battery voltage, all the direct current may be by-passed around the thermo-junction ammeter, so that it will not interfere with the measurement of the alternating current component passing through the circuit. Smaller alternating currents can in this way be measured. Where a large number of measurements are to be made, this circuit involves inconvenience of manipulation.

(3) The third modification of the ammeter-voltmeter method involves the use of a valve voltmeter and an amplifier, but permits of accuracy of measure ment and convenience of manipulation.

12. **The Testing of Audio-frequency Transformers :** F. Benz : *Elektrotech. u. Maschinenb.*, 1928, **46**, 13-17.

This simple article gives frequency-response characteristics for various transformers obtained by measuring the secondary-circuit voltage by means of a valve voltmeter and the primary current by a thermo-junction ammeter. The author suggests that for a comprehensive transformer test, measurements should be made of the resistance and inductance of the windings, the self and mutual capacity, and the step-up ratio at different audio-frequencies. No attempt to obtain general design information is made.

13. **An Optical Method of Measuring Small Vibrations :** H. A. Thomas and G. W. Warren : *Phil. Mag.*, 1928, **5**, 1125-30.

An interferometer method of measuring small mechanical vibrations, such as those encountered in telephone diaphragms is described, and its application

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to the measurement of the vibration of a stiff reed is considered. Substantial agreement with the theory of such a vibration is obtained, and its use is suggested for the study of the vibration of light moving systems. In this case no contact is made and the system is left undisturbed, the only requirement being a small metal surface on the moving part to act as a mirror; a thin metal film could be sputtered on to the surface, as it is not necessary to have a perfectly flat surface. The limiting accuracy is $0{\cdot}5 \times 10^{-6}$ cm., and movements of the order of 10^{-3} mm. can be measured with an accuracy of 0·5 per cent., while movements of about $5{\cdot}10^{-3}$ mm. will have an accuracy of about 0·1 per cent.

14. The Design of Transformers for Audio-frequency Amplifiers with pre-assigned Characteristics : G. KOEHLER: *Proc. Inst. Radio Eng.*, 1928, **16**, 1742-70.

A universal type of bridge is illustrated for making the necessary impedance measurements in connection with the study of transformers. This bridge is adapted to measuring iron-cored coils which carry both alternating and continuous currents, when the reactance is either inductive or capacitative. The bridge can be balanced without distributing the D.C. through the transformer. An amplifier is used as the detecting instrument, and the voltage across the transformer is measured by a valve voltmeter, which latter must be disconnected from the bridge during the final balancing.

The bridge components are not screened, but it is claimed that sufficient accuracy is obtained for most practical purposes.

(See also Abs. 3.3/18.)

15. Low-frequency Intervalve Transformers : R. JOUAUST: *Onde Élect.*, 1928, **7**, 437-45.

After a mathematical treatment of the subject, the author shows that these transformers can only be studied accurately under actual working conditions. A bridge scheme for quantitative measurements is described, and the effects of leakage and virtual grid-filament resistance are considered.

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4.5. DISTORTION IN RECEIVERS

THE three types of distortion which can occur in a radio receiver have already been classified (see 3.4). They are—

(i) Frequency Distortion.
(ii) Amplitude Distortion.
(iii) Wave-form Distortion.

Frequency distortion is the name which has been given to the type of distortion in which different frequencies are amplified unequally. The frequency-distortion which can occur in an audio-frequency amplifier has been discussed in Section 3.4 and is due to the shunt capacity effects introduced by the wiring and the input impedance of the valves in the case of a resistance-capacity coupled amplifier. With transformer coupled amplifiers, distortion is due to the difficulty in designing and manufacturing a transformer which has equal efficiency at all frequencies in the musical range. The design of such transformers has been considered in Section 3.3, where the difficulty is pointed out of securing an adequately large inductance without the attendant self-capacity which a large winding involves. The methods of measuring this type of audio-frequency distortion are fully discussed in Section 4.4.

In the case of a resistance-capacity-coupled radio-frequency amplifier, the effect of shunt capacities is more serious than in the audio-frequency case, but over the limited wave length band representing the side band transmission, the amplification is sensibly constant, and distortion due to this factor is negligible. With a tuned radio-frequency amplifier, however, this is not the case, and such an amplifier in conjunction with a tuned input circuit may produce a type of audio-frequency distortion due to the cut-off of the side band frequencies. The process of modulation of a carrier wave by a series of audio-frequency waves gives rise to the production of a large number of radio-frequency combinations. Each audio-frequency wave produces two radio-frequencies, one greater and one less than the carrier frequency by the amount of the original audio-frequency. If the hollow tone associated with a preponderance of low audio-frequencies is to be avoided, the overtones so necessary for fidelity must be preserved. These overtones are transmitted by the more remote side bands, and if we assume that the modulation includes frequencies up to 10,000 cycles per second, the total width of the transmitted spectrum will be 20 kilocycles. The two side bands are not both necessary for distortionless reception; either may be suppressed at the transmitter without affecting the fidelity of reproduction, and in some transmissions, one side band is completely suppressed, but when both are trans-

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mitted, both must be received in their entirety. The carrier frequency may also be suppressed and be supplied by the receiving station, and, providing the relative amplitudes of the side band frequencies after reception are the same as the relative amplitudes of the initial audio-frequencies, the audio-frequencies after rectification will bear the same amplitude relations to one another as they did before modulation.

The condition of distortionless reception is that all the frequencies in the transmitted band of the modulated wave shall be received in their true relative intensities. This condition is not obtained in practice owing to the resonance characteristics of the transmitting and receiving circuits, and the amount of distortion to be expected from this source can be obtained by superimposing the overall resonance characteristic of the receiver upon a rectangle of width 20 kilo-cycles, or, in the case of single side band transmissions, only 10 kilo-cycles; this applies if the carrier frequency is or is not transmitted. Experimental curves showing the amount of such distortion both at the transmitting and the receiving station are given by SANDEMAN and KIPPING (see Abs. 4.5/6), and show the amount of distortion to be expected. Retro-action at radio frequencies, if not carried to excess, merely serves to modify the impedance of the circuit, but at the same time such radio-frequency retro action will invariably produce a more peaky response characteristic, thus accentuating the frequency distortion. The retro-active effects may include not only those intentionally introduced by inductive or capacitative coupling; they may also arise from valve capacities, leakage and unbalanced wiring capacities, in addition to accidental coil couplings and capacities between coils; screening, of course, assists in preventing such retro-action. Measurements of this effect are few and little detailed information is available. It is clear that single side band transmission is likely to produce better results than double side band transmission, due to the narrower band required to be covered.

The closeness of the present wave length allocations for European broadcast services requires in the receiving circuit, if interference is to be avoided, a degree of selectivity which is incompatible with fidelity of audio-frequency reproduction if ordinary single circuit tuning is used.

The ideal solution of this problem is to use a system which has a substantially rectangular frequency characteristic—that is, one which gives substantially uniform reception over a definite band of frequencies, including all the side band frequencies of the modulated wave, with a cut-off for all frequencies outside this band. The problem permits of two distinct solutions:—

(i) The use of one, or more, band filters, each of which possesses a substantially rectangular frequency characteristic.

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(ii) The use of two, or more, receiving elements whose individual frequency characteristics are not rectangular but which, in combination, have an overall characteristic which is substantially rectangular.

VREELAND (see Abs. 4.2/14) has examined the response characteristics of such circuits by means of the cathode ray oscillograph. The oscillograms show that a substantially rectangular characteristic can be obtained by the use of three resonance circuits, the tuning points of which are slightly dissimilar. With only two tuned circuits the characteristic has a double hump and is rather similar to that obtained with two coils tightly coupled.

Amplitude distortion in audio-frequency amplifiers has been discussed in Section 3.4. Its measurement presents no difficulties, as the ordinary measuring apparatus used for the determination of amplification may be utilised, providing that both the input and the output are separately measured. In well designed audio-frequency amplifiers, in which amplitude distortion is produced only by overloading the valves or transformers, this type of distortion should be small. Amplitude distortion arising in radio-frequency amplifiers is more serious, particularly when several stages are used. Some curious effects are stated to have been observed, the explanation of which is not clear. An examination of the relation between input and output for various types of radio-frequency amplifiers is desirable.

Wave form distortion occurs when the amplification process gives rise to frequencies which are not necessarily present in the original input wave form. The causes of such distortion in audio-frequency amplifiers have been discussed in Section 3.4. In radio-frequency amplifiers, wave form distortion can occur as a result of:

(i) Operating the valves beyond their normal linear dynamic characteristics.

(ii) Permitting the grid swing to make the grid positive during part of the periodic cycle, thus introducing grid current.

(iii) Grid-condenser charging effects, which give rise to transient distortions when excessively large peak voltages occur.

Such distortion will give rise to radio-frequency harmonics of the carrier wave, and, after detection, these harmonic components will create audio frequencies at multiples of the original modulation frequency. Usually these double and treble modulation frequency terms are small, but if serious radio-frequency distortion occurs, perceptible audio-frequency effects may be introduced.

Rectification with an ideal detector should reproduce the original modulation as an identical audio-frequency output, but in the case of actual rectifiers a number of additional frequencies, not present in the original audio-frequency wave are introduced; these

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are sometimes spoken of as combination tones. This fundamental property of the rectifier is fully discussed in Section 2.1.

The measurement of such distortion involves the determination of the output wave form, for which purpose an oscillograph is required. Four instruments are available for this purpose :—

(i) The Duddell Oscillograph.
(ii) The Einthoven Galvanometer.
(iii) The String Electrometer.
(iv) The Cathode Ray Oscillograph.

The Duddell Oscillograph requires a reasonably large power input and has an upper frequency limitation of about 5,000 p.p.s. The method has been found useful in determining the permissible undistorted output which can be obtained from various power output stages (see Abs. 4.5/10).

The Einthoven Galvanometer requires less power to operate it. It has been used by THOMAS (see also 4.5/7) for the examination of the output wave form delivered by small power output valves, but the upper frequency limitation, about 800 p.p.s., limits its utility.

The String Electrometer possesses the advantage of requiring little power, but the same limitations of frequency are imposed due to the inertia of the moving system.

The Cathode-Ray Oscillograph is eminently suitable for the examination of wave-form owing to the almost instantaneous response and the almost complete lack of inertia of the moving system. High voltage oscillographs, such as the Dufour, are unsuitable for amplifier work owing to their poor sensitivity, but the low voltage type has been used with considerable success. The best known oscillograph of this type is that made by the Western Electric Co.

For the purpose in hand, an oscillographic method of recording the wave form may be employed. The distinguishing feature of this method is that one motion of the cathode beam is controlled by the unknown wave, while the second component motion has as its sole function the provision of a time scale or time co-ordinate, the purpose of this on the indication or record being to render possible the study of the time variations of the electrical quantity under investigation.

The simplest method consists in applying a sinusoidal wave to one pair of deflecting plates and the unknown wave to the other. If the two frequencies are the same, a Lissajou figure is obtained which can be analysed in the usual manner. The difficulty lies in guaranteeing the perfection of the sinusoidal wave.

DYE (see Abs. 4.5/3) uses either a circular, or an elliptical, time scale and superimposes the unknown wave form upon this base. A low audio-frequency source supplies current to a resistance and a condenser in series, the voltages across which will, therefore,

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be in quadrature. By connecting the resistance to one pair of deflecting plates and the condenser to the other pair, a deflection is produced in which the ratio of the major and the minor axes is determined solely by the ratio of the resistance to the reactance of the condenser. The potential source under examination is inserted in series with that pair of deflecting plates producing the minor axis, and the frequency of the time base is adjusted to be a sub-multiple of the frequency of the wave under examination. For periodic variations, the oscillogram takes the form of a stationary pattern when the two frequencies are synchronised. By adopting a small minor axis, a portion of the pattern is given with very nearly a linear time base and the wave form extracted from this portion of the pattern can be used for the delineation of the harmonic constituents of the wave form. This method, however, has been superseded by recurrent linear time base methods, but has advantages in frequency measurement work.

Various methods of imposing a linear time base have been produced; by this means the unknown wave form is plotted on rectangular co-ordinates. The "blinking" Neon lamp method developed by KIPPING (see Abs. 4.5/2) makes use of the potential difference across a resistance in series with a Neon lamp circuit. During the greater part of the time period (while the lamp is out) the current flowing into a condenser decreases according to an exponential law, but with a suitable choice of the constants of the circuit and the characteristics of the Neon lamp, it may be made very nearly a linear variation, so that the motion of the beam may be considered as a periodic uniform motion. The lighting of the lamp causes a large rise in the current, giving a very rapid return motion to the beam.

In ROGOWSKI'S method (see Abs. 4.5/5) the electric deflection of the beam is controlled by the potential difference across a condenser, and this potential difference is given a triangular wave form by making the current flow through the condenser of rectangular wave form, i.e., the condenser is charged with a constant current and discharged with a constant current, this result being secured by two opposed diodes working on their saturation current. The practical difficulties of realising this condition of rectangular current wave form were investigated by ROGOWSKI and GLAGE. ROGOWSKI overcame these difficulties by using unequal heating of the two diode filaments. By this means the discharge of the condenser could be accomplished with a higher current and, therefore, in a shorter time than the charge; a saw-tooth diagram is thus obtained which prevents overlapping of the wave shapes. An alternative method tried by him is to allow the condenser to discharge through a spark gap instead of through the second diode. This arrangement simplifies the apparatus required, but there may be complications due to radio-frequency oscillations accompanying the discharge of the condenser.

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Perhaps the best method of this type is that due to APPLETON, WATSON WATT, and HERD (see Abs. 4.5/8). In this method the time base plates are connected across the grid condenser of a simple radio-frequency oscillator using inductive coupling, in which the usual grid leak is replaced by a saturated diode. By this means the condenser discharge is modified from its usual exponential form to a linear form, and the wave form under analysis is plotted with rectangular co-ordinates. A complete portable equipment embodying this principle has proved useful for the delineation of wave forms at the NATIONAL PHYSICAL LABORATORY (see Abs. 4.5/8).

It must be remembered that the accuracy with which a wave form can be analysed by means of any cathode ray oscillographic method is limited by the dimensions of the spot on the fluorescent screen. To obtain deflectional sensitivity, low-velocity beams must be used. Accompanying the use of these, however, there are two serious disadvantages: firstly, the decrease in the recording ability of the beam due to its lower velocity; secondly, the increased "scattering" tendency of the beam, which not only destroys the accuracy of the instrument but also detracts further from the recording ability. The requirement here is obviously an improved method of focusing the beam. Unless such a focusing method of itself gives sufficient improvement, it will be necessary to devise independent means for increasing the recording sensitivity of the instrument. One method that presents itself is to increase the power in the beam available at the recording screen without, if possible, affecting its deflectional sensitivity. Alternatively, the fluorescent layer may be improved to render it more sensitive to the impact of slow electrons—for example, by converting a higher proportion of the incident energy into light energy. Finality has not been reached in the design of cathode ray oscillographs, and it is hoped that improvements in the high sensitivity oscillograph may be made, so that the analysis of wave-form distortion may be extended to cases which cannot as yet be examined owing to the limitations imposed by the measuring apparatus.

Instead of using an oscillograph and analysing the wave form into its constituent harmonic components, the amplitude of each harmonic component may be measured directly by means of a tuned system which picks out the successive harmonics. If the current in a resonant circuit is recorded while the tuning of the circuit is continuously varied, the curve connecting the current with the tuning frequency will show peaks when the system is tuned to any harmonic frequency of the input. By comparing the value of these peaks, it is possible to find the harmonic constituents of the original wave. A knowledge of the effective impedance of the resonance circuit at each frequency is of course required. A highly complicated instrument has been devised by WEGEL and MOORE (see Abs. 4.5/4) which operates on this principle, the final result being given in the form of a curve in which the relative amplitudes of the

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peaks correspond to the energy existing in each harmonic constituent of the original wave. It is stated that the amount of power required does not exceed in general 500 microwatts, and the frequency range is from 20 to 5,000 cycles; the time necessary to take the record is about five minutes. This apparatus has been satisfactorily used for the analysis of the audio-frequency output given by an amplifier, but the apparatus involved is of a highly complicated nature and is too expensive for most laboratory purposes, unless a very large number of waves require to be analysed, in which case the saving of time due to the use of this instrument would be a justification for its complication.

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1. **The Unilateral Dynamic Characteristics of Three-electrode Vacuum Tubes :** J. C. FRAYNE : *Phys. Rev.*, 1922, **19**, 629-51.

(See Abs. 3.4/1.)

2. **A Practical Demonstration of some Applications of the Cathode Ray Oscillograph :** N. V. KIPPING : *Wireless Wld.*, 1924, **13**, 705-9.

The author describes various applications of the cathode-ray oscillograph to the solution of problems existing in audio-frequency amplifier circuits; particular attention is given to the Neon lamp method, in which a Neon lamp supplied from a direct current source, and shunted by a variable condenser, is made to " blink " at a sub-multiple of the frequency of the wave form to be examined. The type of time base given by this arrangement is approximately of a saw-tooth shape; since the time of charge and discharge is about the same, a double trace is obtained, the super-position of which give the final pattern on the screen. There is, of course, no difficulty in discriminating between the two and arriving at the true wave form.

3. **Improved Cathode-Ray Tube Method for the Harmonic Comparison of Frequencies :** D. W. DYE : *Proc. Phys. Soc. Lond.*, 1924, **37**, 158-68.

A portion of this paper deals with a method of examining audio-frequency wave-forms. The cathode beam is given an electrical movement at a low audio-frequency by means of a two-phase supply, obtained in the following manner; an audio-frequency generator supplies current to a resistance and a condenser in series, the voltages across which will therefore be in quadrature; by connecting the resistance to one pair of deflecting plates and the condenser to the other pair, an elliptical deflection is produced in which the ratio for the major to the minor axes is determined solely by the ratio of the resistance to the reactance of the condenser.

The potential source under examination is inserted in series with that pair of deflecting plates producing the minor axis, and the frequency of the time base is adjusted to be a sub-multiple of the frequency of the wave under examination. For periodic variations, the oscillogram takes the form of a stationary pattern when the two frequencies are synchronised. By adopting a small minor axis, a portion of the pattern is given with very nearly a linear time base, and the wave form extracted from this portion of the pattern can be used for the delineation of the harmonic constituents of the wave form.

Some examples are given in the form of reproductions from prints made directly on pieces of bromide paper held down on the top of the cathode-ray tube for a few seconds.

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4. **An Electrical Frequency Analyser :** R. L. WEGEL and C. R. MOORE : *Bell System Techn. J.*, 1924, **3**, 299-323.

An apparatus has been developed by means of which it is possible to measure and obtain a permanent record of the frequency components of an electric current wave. The device has two frequency ranges ; 20 to 1,250 cycles per sec. and 80 to 5,000 cycles per sec. ; the amount of power required does not in general exceed 500 micro-watts, and the time necessary for making a record is about five minutes. An attachment is provided which permits the making of simultaneous harmonic analyses of two complex waves at the same time.

In principle, the process consists in introducing the complex wave to be analysed into a selective network, the essential feature of which is a sharply tuned circuit, the resonant frequency of which is controlled by varying the capacity in small steps by means of a pneumatic apparatus similar to that in a piano-player. A maximum response of the circuit occurs at each frequency of tuning which coincides with a component of the complex wave. An automatic photographic recorder of the response to each frequency of tuning is provided, by means of which the frequency and magnitude of each component of the complex wave may be obtained. For convenience of operation, an automatic control apparatus is provided, so that it is only necessary to connect the complex source or sources to be analysed and press a starting button. The completed record of the analysis is delivered after the machine has passed through the entire range of frequencies.

The apparatus has been used for the analysis of the distortion which may exist in audio-frequency amplifiers and oscillators.

5. **Measurements in Electrical Engineering by means of Cathode Rays :** J. T. MACGREGOR-MORRIS and R. MINES : *J. Instn. Elect. Eng.*, 1925, **63**, 1056-1107.

The paper is a comprehensive survey of the present state of knowledge on the subject of the use of cathode rays for measurement purposes in electrical engineering, and specially with a view to future developments. Part I introduces the subject of measurement methods. A description is given of the various ways in which information on physical quantities may be obtained by applying them to suitably devised instruments, together with the history of the development of such instruments. Various methods of producing cathode rays are given and the manner in which these are applied to measuring instruments is indicated. Part II deals in greater detail with measuring apparatus using cathode rays. The historical development of such apparatus is outlined and references are given to the purposes for which it was devised. Certain special problems arise in the use of such apparatus for accurate measurement and an indication is given as to how these problems have been solved. A description is also given of the auxiliary apparatus necessary for the study of electrical oscillations by means of cathode rays. Part III contains a brief review of the limitations of the methods described in the paper, and indicates some possible lines of future development. An appendix gives a mathematical investigation of certain essential properties of the cathode beam as used in the apparatus described.

6. **Distortion in Radio-telephony :** E. K. SANDEMAN and N. KIPPING : *Experimental Wireless*, 1925, **2**, 756-64, 811-19.

This article discusses firstly the various types and causes of distortion and, secondly, the application of the cathode-ray oscillograph to the investigation of distortion. The authors state that a sensibly uniform amplification-frequency characteristic can be obtained by means of tuned transformer audio-frequency amplifiers.

Amplitude distortion may occur if any part of the apparatus is worked on non-linear portions of its characteristic. The rectifier will necessarily

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involve some degree of distortion for this reason. Grid-current detection will also give amplitude distortion owing to the variation of the audio-frequency impedance of the resistance-condenser combination.

The various ways of using a cathode-ray oscillograph to measure wave-form distortion are described, together with the use of the oscillograph for the measurement of the percentage of modulation. The paper concludes with a description of the Neon lamp method of producing a linear time base; this is given in the form of an Appendix.

7. **The Performance of Amplifiers:** H. A. THOMAS: *J. Instn. Elect. Eng.*, 1926, **64**, 253-68, Disc. 268-78; *Experimental Wireless*, 1926, **3**, 15-18; *Elect. Rev. Lond.*, 1926, **98**, 73-4

The portion of this paper applicable to this section deals with a method of examining the output wave form at frequencies below 300 p.p.s. by the use of an Einthoven galvanometer, and also at frequencies above 300 p.p.s. by the use of a special cathode-ray oscillograph equipment using a linear time base. Experimental results obtained with these instruments are given, and the magnitudes of the harmonics introduced with various arrangements of the amplifier are tabulated. It is shown that iron saturation in the inter-valve transformers is responsible for a considerable amount of wave-form distortion.

(See also Abs. 1.2.7/6, 1.4.1/16, 3.4/9, 4.2/8, 4.4.1/7, 4.6/1.)

8. **The Delineation of Alternating Current Wave Forms:** H. A. THOMAS: *Experimental Wireless*, 1927, **4**, 15-23.

For the study of wave-form distortion at low audio frequencies below about 300 p.p.s. an Einthoven String Galvanometer may be used. By inserting this instrument in the anode circuit of the last stage of amplification, a photographic record of the wave form can be obtained. For higher frequencies, the cathode-ray oscillograph must be used. By connecting one pair of deflecting plates across the output terminals, and the other pair across a tuned coil loosely coupled to the input oscillator, a Lissajous figure is obtained which can be photographed and analysed. This resolution is tedious, and a better method is to employ a synchronous linear time base, in which case the wave form is depicted directly with rectangular co-ordinates. The various methods of obtaining such a linear time base are mentioned, and a full description of the method finally adopted is given. This method is due to APPLETON, WATSON WATT, and HERD, and consists essentially in connecting the time-base plates across the grid condenser of a simple radio-frequency oscillator using inductive coupling, in which the usual grid leak is replaced by a saturated diode. By this means, the condenser discharge is modified from its usual exponential form to a linear form. The complete apparatus is illustrated in detail.

9. **Load Carrying Capacity of Amplifiers:** L. MELHUISH and F. C. WILLIS: *Bell System Techn. J.*, 1926, **5**, 573-92.

The use of a cathode ray oscillograph for the determination of wave-form distortion in amplifiers is described in detail. The sinusoidal input voltage produces a horizontal deflection of the cathode beam, while the output voltage produces the vertical deflection; in this manner, the distortion which is present is readily seen by noting the resulting figure on the fluorescent screen. If the figure is a straight line or an undistorted ellipse, the amplifier is not introducing distortion, but when distortion occurs, the oscillograph shows a sharp bend at one or both extremities of the line, or an obvious distortion of the ellipse.

Although measurements with such an oscillograph are limited in accuracy by the imperfections of the oscillograph, the simplicity of the

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method establishes it as a convenient means of measuring wave-form distortion. The accuracy obtained by the use of an oscillograph is probably sufficient for most practical purposes, although, if greater refinement is required, the harmonics can be separately measured by systems of tuned circuits; this latter method is very complicated.

10. **An Examination of the Properties of Audio-frequency Amplifiers by means of the Duddell Oscillograph**: W. BAGGALLY: *Experimental Wireless*, 1926, **3**, 628-32.

The author shows that average general purpose valves are usually overloaded on medium loud-speaker outputs, and that large power valves are required to give reasonably large undistorted volume.

The method adopted in the measurements consisted in substituting the Duddell Oscillograph for the loud-speaker and applying a pure sinusoidal wave at a frequency of 480 p.p.s.

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4.6. THE OVERALL PERFORMANCE OF RECEIVERS

THE aim of all measurements of the constituent parts of a radio receiver is ultimately to obtain a characteristic which represents the behaviour of the complete receiver. Such a measurement is the most important of all the measurements which can be performed and at the same time is the most difficult. The difficulty lies in the fact that the output quantities are in general very large compared to the input quantities, so that the slightest form of coupling between the input and the output circuits introduced by the measuring apparatus may completely modify the behaviour of the receiver. Another difficulty lies in the fact that the input and output frequencies are dissimilar.

All early attempts to obtain the overall characteristics of a receiver made use of the assumption that the overall performance could be represented in terms of the product of the performances of the separate components. If this were true, the problem would be simple, but unfortunately the effects due to the input impedance of an amplification stage, and due to various forms of retro-action, render this assumption unjustifiable in all but the very simplest cases.

Due to the difference in frequency between the input and output, the large magnitude of the step-up ratio, and the retro-active effects, no form of substitution, bridge, or comparison method is admissible. The input must be supplied with an artificially generated signal, the form of which is identical with that of the actual signals with which the receiver is normally supplied in practice, and the relatively large output must be measured by a method which in no way disturbs the normal operating function of the receiver.

An overall performance measurement implies the inclusion of the input tuned circuit, the antenna or the frame, whichever is used in practice with the receiver. The problem of applying a known E.M.F. to this tuned circuit may be solved in the following ways :—

(i) By inducing the E.M.F. into the main tuning inductance by permitting this inductance to act as the secondary circuit of a known mutual, the current in the primary of which is known.

(ii) By passing a known current through a condenser of known value which forms part of the tuning capacity.

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(iii) By inserting a small non-inductive resistance or a small known inductance in series with the tuned circuit carrying a known current, by which means a known E.M.F. is applied.

(iv) By using a current transformer and potentiometer such as a Dye transformer, or an attenuator having known properties.

The first method has been used but involves the measurement of a mutual inductance of large ratio at radio-frequencies (see Abs. 4.6/3, 5). It is impossible to obtain such information with any accuracy and the method is unsatisfactory from this point of view. Another difficulty of this method lies in the fact that the tuning circuit of the receiver must always include, or consist of, this secondary circuit of the calibrated mutual. The method is, therefore, inapplicable as a general method of testing receivers in which the tuned circuit forms part of the complete assemblage.

In the second method, since the value of the tuning condenser is specified by the wave-length range, the E.M.F. which can be applied can only be controlled by the current passed through this condenser and supplied by the main oscillator. Consequently the possible range of the input E.M.F. is very limited. For the smallest measureable radio-frequency current, the value of this condenser must be very large if the E.M.F. across it is to be of the right order.

The third method has been adopted to a greater extent than the previous methods, owing to its more general application and its comparative independence of frequency. Small non-inductive resistances can be made within fairly wide limits, and together with a reasonable range of current, the method is applicable to most practical cases. This method has been adopted by THOMAS, OAKLEY and HULL (see Abs. 4.6/1, 6 and 4.2/9).

This method, which has yielded the most consistent results, adopts a potentiometric input into the tuned circuit, and a measurement of the output by means of a valve voltmeter. A radio-frequency oscillation modulated to a known extent at an audio frequency to a known extent, supplies the input potentiometer. The prevention of direct magnetic and electric induction from the oscillators into the amplifier demands very careful screening of these components, and of all leads up to and including the potentiometer itself. This is effected by THOMAS, OAKLEY and VAN DYCK (see Abs. 4.6/7) by enclosing the oscillators in a completely screened metal room. The amplifier, too, must be housed in a screened room to prevent induction from external sources. Refinement in measurement is dependent upon the perfection of these screens and nothing much short of complete shielding is permissible.

The potentiometer takes the form of a straight resistance wire with a concentric return, by which means complete shielding is effected. If the resistance wire is tapped at various points, E.M.Fs. of known magnitudes may be obtained. At very high frequencies

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it is important that the current through the wire should be measured at a point as near to the tapping as possible, from which it appears that Oakley's apparatus has an upper frequency limitation, due to the fact that the thermo-junction ammeter is situated in the oscillator screened room. If the filament of the amplifier is connected to the junction of the potentiometer-resistance wire and the concentric return, the batteries will be virtually at earth potential, which condition is required for the normal operation of the receiver. It is important that this battery-earthing connection is adhered to, otherwise the capacities to earth may modify the measured performance. If the amplification is very large, the length of even a very short resistance wire becomes too great, and it is then necessary to replace this wire by a copper conductor, whose inductance per unit length can be calculated. HULL (see Abs. 4.2/9) has succeeded in making use of this arrangement for amplifications up to 2,000,000 at a frequency of 1,000 kilo-cycles per sec.

The fourth method makes use of some means of reducing the current applied to the input potentiometer by a known amount. Dye's current transformer (see Abs. 4.2/7) is very suitable for this purpose.

Several methods employ an " attenuator," or system of circuits, in which the output current is a definite and known ratio of the measured input current (see Abs. 4.6/4, 5, 7).

An examination of the methods of measurement which have been adopted indicate that the last two methods are the ones most likely to yield satisfactory results in future work. The factors limiting the accuracy in practice are :—

(i) The impossibility of measuring the effective impedance of the input potentiometer, and the consequent dependence of this value upon theory. However, the possibilities of error in this determination are small.

(ii) The difficulty of measuring the current existing in the potentiometer at very high radio-frequencies. At present a non-contact, thermo-junction ammeter is used, and the assumption is made that its direct current calibration is valid at all frequencies. A better method of measuring very high radio-frequency currents would naturally improve the accuracy of measurements at such frequencies.

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This paper describes the methods and apparatus used in the early stages of the development of the National Physical Laboratory method of measuring the performance of amplifiers. The modulated input was supplied to the

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receiver by means of potentiometers and step-down transformers, and the output was measured by means of a vibration galvanometer calibrated by means of the modulator. A device was used for quickly determining the condition of a 100 per cent. modulation. If a tuned circuit and crystal rectifier with telephone output is coupled to the radio-frequency source, when the modulating percentage exceeds 100, a second harmonic of the modulator fundamental is introduced into the rectifier output, and this is readily detectable by the ear. The accuracy with which the condition of complete modulation may be obtained by this method is about 5 per cent., which is sufficient for the purpose of rough measurements. The elaborate screening arrangements which were found to be necessary are described in detail, some results of overall performance measurements are given, and the general properties of some typical receivers are discussed.

(See also Abs. 1.2.7/6, 1.4.1/16, 3.4/9, 4.2/8, 4.4.1/7, 4.5/7.)

2. **Importance of Laboratory Measurements in the Design of Radio-receivers :** W. A. McDONALD : *Proc. Inst. Radio Eng.*, 1927, **15**, 99-111.

The author stresses the importance of the accurate measurement of certain quantities in relation to the design and production of commercial broadcast receivers. A list is given of the thirteen measurements which are considered necessary to specify the properties of the receiver, and the equipment for making these measurements is described.

The author falls into the error of assuming that the individual properties can be separately analysed, and the final overall performance specified in terms of the separate properties.

The methods of measurement lack refinement, and no precautions are taken to prevent the valve voltmeter load seriously affecting the amplification ; the author uses the valve voltmeter in such a manner that one must conclude that he has failed to realise the significance of the loading effect of this instrument.

3. **Notes on Radio-receiver Measurements :** T. A. SMITH and G. RODWIN : *Proc. Inst. Radio Eng.*, 1927, **15**, 387-95.

The authors choose the characteristics of sensitivity, selectivity and fidelity as the most important for measuring the performance of a receiver and, in the tests, take a standard loud-speaker signal as that corresponding to an average audio-frequency voltage of 15 across a resistance of 5,000 ohms inserted in the output circuit, when 50 per cent., 400 cycle modulation is furnished. Typical curves are shown for (*a*) sensitivity, plotted between the field strength required to produce the standard signal and frequency ; (*b*) selectivity, given by a resonance curve plotted between the field strength required to produce the standard signal and the frequency of the signal generator ; and (*c*) fidelity, by combining curves showing the side-band transmission and the audio-amplifier characteristics.

The input E.M.F. is obtained by induction from the screened modulated radio-frequency oscillator, and therefore the measurement involves the knowledge of the mutual inductance between the dummy antenna and the inducing coil. The difficulty of measuring such a mutual inductance at radio-frequencies is well known.

4. **Apparatus for Amplification Measurements :** F. GABRIEL : *Jahrb. drahtl. Telegr.*, 1927, **30**, 95-100, 123-6.

The apparatus described has been developed for the measurement of the overall amplification of amplifiers or multiple valves. The input is supplied by a transformer, in which the primary current is measured by a thermo-junction ammeter. A large range of inputs is obtained by the use of shunts

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and series resistances. Such an arrangement is unsuitable for high radio-frequencies. The output is measured by a thermo-junction ammeter in the anode circuit of the last stage.

No screening precautions appear to have been taken.

5. **The Quantitative Measurement of Receivers:** E. KLOTZ: *Telefunkenztg.*, 1927, No. 50, **9**, 34-38.

The paper describes apparatus developed and employed in the Telefunken Laboratories for testing radio receivers. A modulated radio-frequency source is used to supply a potentiometer, a tapping from which is taken to the input terminals of the receiver. A mutual inductance is used between the potentiometer and the main oscillator, the coils of this mutual inductance being electrically screened from each other. It is claimed that amplification factors as high as 3×10^6 can be measured, and in this case the input is of the order of 0·005 microvolt. The valve voltmeter used to measure the output has a range of from 0·02 to 100 volts, and the limiting frequency which can be used is 3×10^6.

Sufficient care does not appear to have been taken in the design of the potentiometer, and the screening arrangements are insufficient.

6. **Quantitative Determination of Radio Receiver Performance:** H. D. OAKLEY: *J. Amer. Inst. Elect. Eng.*, 1927, **46**, 568-72.

The author points out the impossibility of obtaining the overall performance of a receiver from computations based upon the measured performance of the individual sections. He then proceeds to describe the apparatus used by the General Electric Company to obtain the overall characteristics of a complete receiver.

A radio-frequency source is modulated according to the Heising scheme, and the output is supplied to an "attenuator" which gives a known small E.M.F. to the input of the amplifier. The output is measured by means of a valve voltmeter connected across the loud speaker terminals. The oscillators, together with the measuring apparatus, are housed in a completely screened room, while the amplifier under test, together with the attenuator and the valve voltmeter, are housed in a second adjacent screened room. The function of the first screen is to prevent direct induction from the oscillators into the amplifier, while the function of the second screen is to screen the amplifier from fields due to external sources. The attenuator consists of a copper wire with concentric return, the point of contact between the two being earthed and connected to the filament of the amplifier. The inductance per unit length is calculable and, by means of tappings, known small voltages can be applied.

The measuring thermal instrument is situated in the oscillator screened room and, therefore, at very high radio-frequencies its reading does not give a true measure of the current in the attenuator.

The input may be applied to the amplifier directly, to a dummy antenna having constants similar to those of an ordinary antenna, or to a frame coil. The results obtained are expressed as audio-output voltage existing across the loud speaker terminals with respect to the known input radio-frequency voltage modulated by a definite percentage. The selectivity of the receiver is expressed in terms of curves connecting the input voltage with definite modulation with respect to the frequency change about the normal resonance point for a constant output across the loud speaker terminals. The properties of the audio-frequency amplifier are determined by varying the frequency of the modulator while maintaining a constant input voltage and a constant percentage of modulation.

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Overall measurements supply the means for obtaining a picture of the performance of a receiver as a whole, and with such pictures the performance of different receivers may be readily compared.

7. **Quantitative Methods used in Tests of Broadcasting Receiving Sets :** A. F. VAN DYCK and E. T. DICKEY : *Proc. Inst. Radio Eng.*, 1928, **16**, 1507-31.

A comprehensive description of apparatus found necessary for the complete testing of broadcast receivers. The testing equipment is of two kinds ; that for laboratory development and research, and that for factory production. In the former, special oscillators are required. In audio-frequency work, a heterodyne oscillator with an output up to 50 volts is used, while for radio-frequency selectivity tests a carefully screened and neutralised oscillator-amplifier is essential. In making overall tests, the radio-frequency currents are modulated to a known depth and the loud-speaker voltage measured by the slide-back method. In production, testing accuracy is sacrificed to robustness and convenience. Sensitivity is measured by means of a modulated radio-frequency oscillator, and a fidelity test comprises modulation by gramophone records. The layout of the apparatus and the screened testing cabins are described in detail and illustrated.

The general arrangement follows usual practice, the audio oscillation being fed into the anode circuit of the radio-frequency oscillator. Both oscillators are elaborately screened and the receiver is situated in a screened cabin. The authors state that the accuracy of reproduction of such measurements is about 15 per cent.

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DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH

RADIO RESEARCH

SPECIAL REPORT No. 10

THE ORFORDNESS ROTATING BEACON AND MARINE NAVIGATION

BY

R. L. SMITH-ROSE, D.Sc., PH.D., A.M.I.E.E.

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PREFATORY NOTE

A PREVIOUS report (Radio Research Special Report No. 6) describes a lengthy series of experiments carried out at Gosport by the Radio Research Board on a rotating loop beacon of the type originated by the Air Ministry for aerial navigation. These experiments indicated that the rotating loop beacon system should form a useful and reliable aid to marine navigation, and, accordingly, the Wireless Direction-Finding Committee of the Board of Trade recommended that an experimental, but more permanent, beacon of similar type should be erected at a place where an extended trial of this system could be made by ships at sea. The site selected was Orfordness, and the cost of the installation and operation of the beacon erected there has been shared equally by the General Lighthouse Fund and the Air Ministry. The chief object of the trial of the Orfordness beacon was to ascertain in particular its performance when observations were carried out by the personnel of various classes of ships of the Mercantile Marine. The results of the observations on the beacon have been analysed on behalf of the Board of Trade by the Radio Research Board of the Department of Scientific and Industrial Research and the present report contains a summary and a discussion of these results.

A. G. LEE,
Chairman,
Radio Research Board.

DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH,
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THE ORFORDNESS ROTATING BEACON AND MARINE NAVIGATION

1. Introductory

During the past ten years the application of directional wireless methods as aids to marine navigation has become very widespread. Prior to 1920 the methods generally adopted consisted in the use of wireless direction-finders at special shore stations erected for the purpose: these stations on request supplied bearings to ships, which made special transmissions for the purpose.

The alternative scheme which has received the most attention since the above date consists in placing the radio direction-finding instrument on board the ship itself, so that radio bearings may be taken upon any transmitting station, the position of which is known to the operator or navigator. The whole tendency of the development of direction-finding by wireless of recent years has been in the direction of enabling the navigator to determine his bearings without reference to the personnel operating any shore station.

As an alternative to the use of a direction-finder on board ship, the rotating beacon transmitter has provided a means of obtaining bearings by wireless without the necessity of any special apparatus on the ship itself, and without the necessity of special signalling between the ship and any shore station. The development of this system of wireless direction-finding was begun several years ago by the Air Ministry in this country to avoid the necessity of carrying any special gear in aircraft navigating by wireless.

The principles and historical development of this rotating beacon have been described in published papers,* the second of which referred in detail to the carrying out of a series of experiments for the Radio Research Board with the object of determining the applicability of this system to marine navigation. The present report deals with a further stage in the progress of this subject, following upon the installation of a rotating beacon transmitter at Orfordness, Suffolk, for the general use of the Mercantile Marine. This beacon was erected by the Air Ministry on behalf of the Board of Trade, half the cost being defrayed from the General Lighthouse Fund.

2. Description of the Principle of the Beacon

The beacon consists of a rotating loop transmitter having directional properties by which an observer in a ship can obtain his bearing from the beacon without the use of special wireless direction-finding equipment in the ship. The Orfordness beacon situated on

* Gill, J. H. and Hecht, N. F. Rotating-Loop Radio Transmitters and their application to Direction-Finding and Navigation. *J. Inst. Elect. Eng. Lond.* 1928, **66**, 241.

R. L. Smith-Rose and S. R. Chapman. Some experiments on the application of the Rotating Beacon Transmitter to Marine Navigation. *J. Inst Elect. Eng. Lond.* 1928, **66**, 256.

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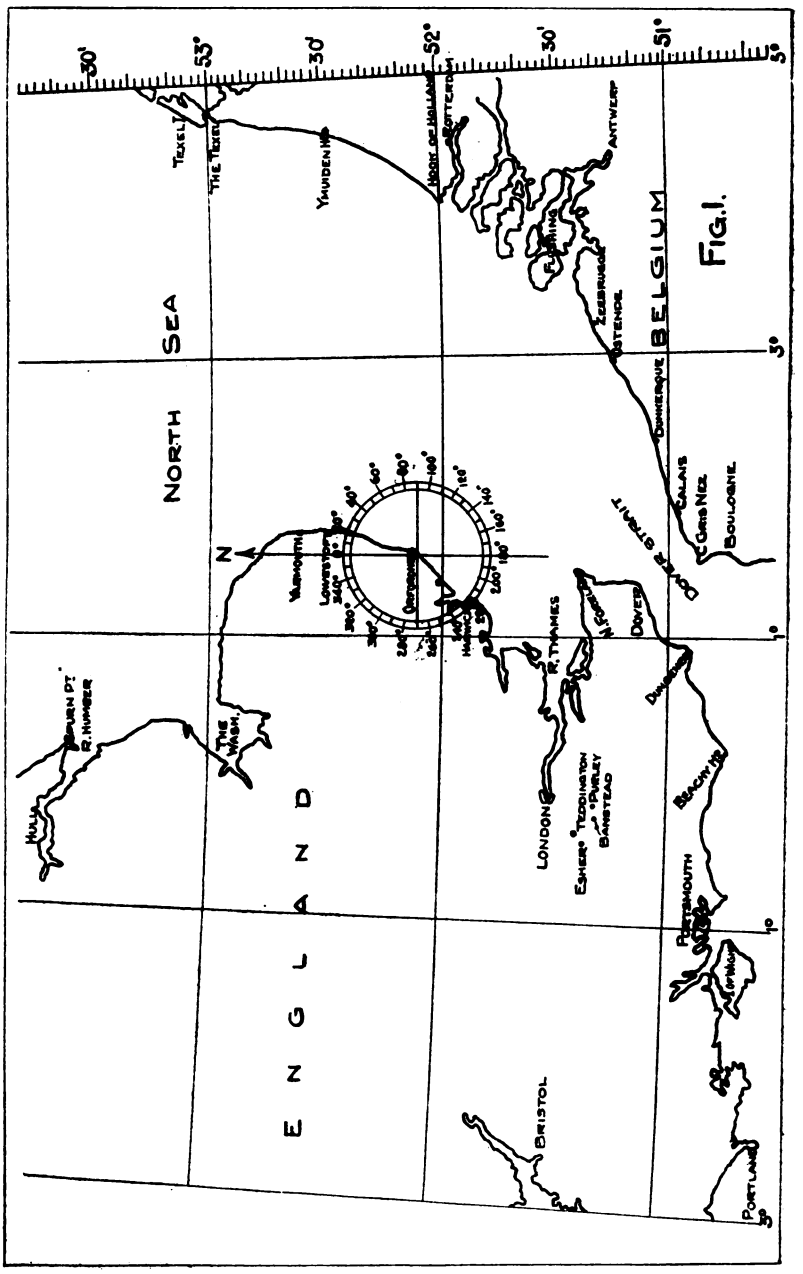

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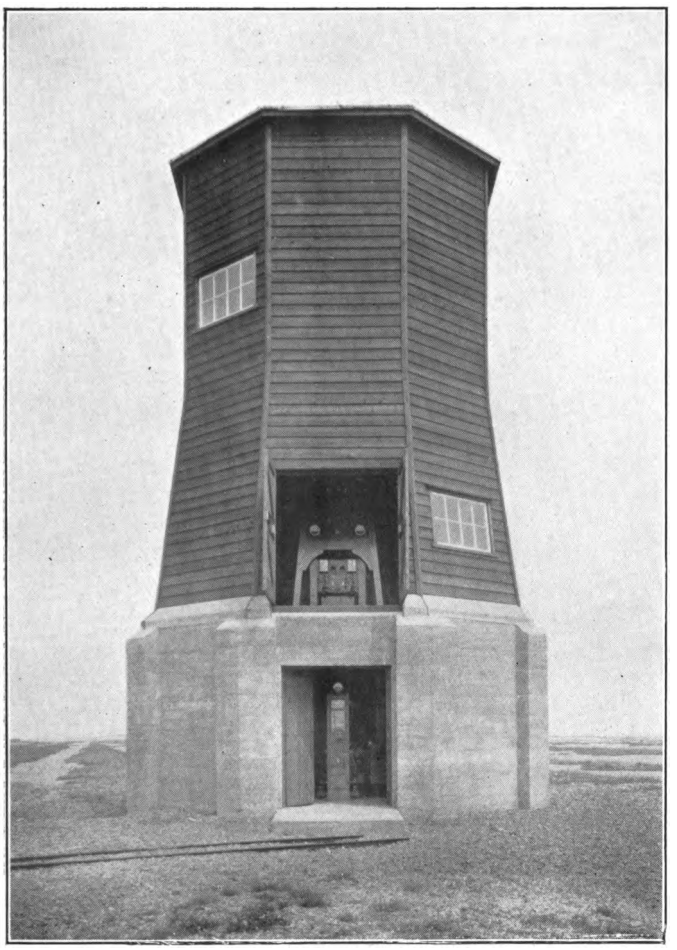

FIG. 2.—*Exterior view of the Orfordness Rotating Beacon station. The rotating loop transmitter with all necessary electrical supply and control gear is completely enclosed in the building shown.*

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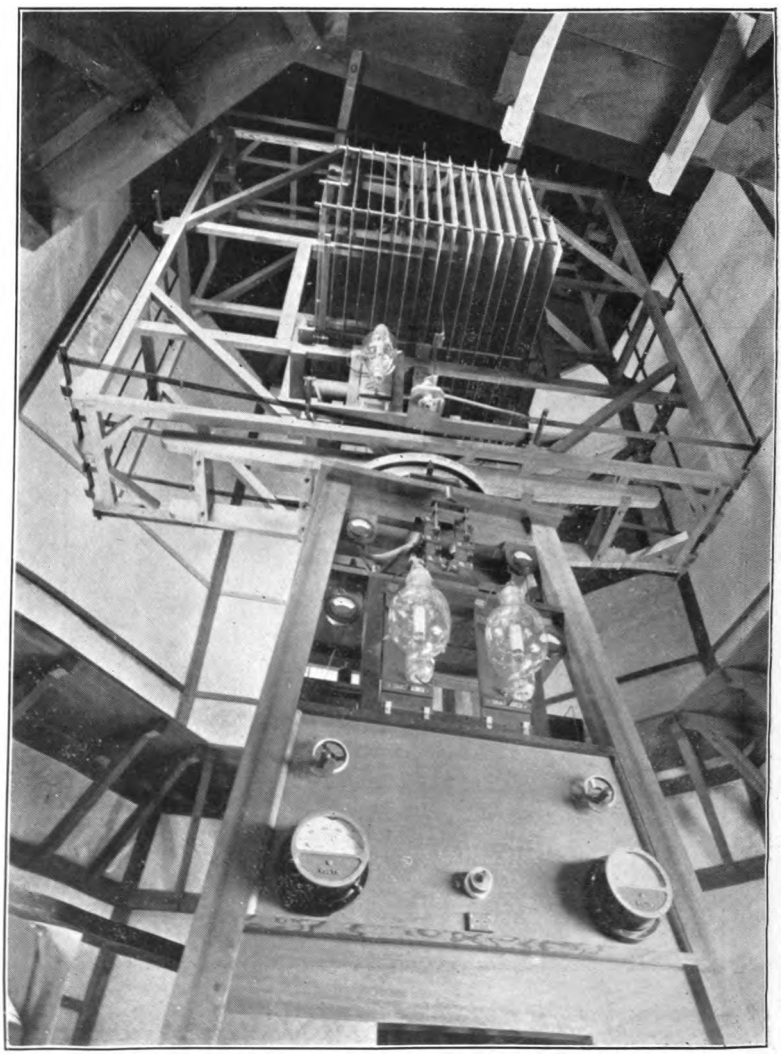

FIG. 3.—*Interior view of the Orfordness Rotating Beacon, showing rotating loop containing transmitting valve, tuning condenser, etc., and the rectifying valves and instruments mounted on the control panel below the loop.*

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the East coast (Fig. 1) operates on a wave-length of 1,040 metres (frequency 288·5 kilocycles per second), and uses the call-sign GFP. The beacon is now in continuous daily operation: each transmission lasts for five minutes, after which the beacon remains silent for five minutes and then automatically starts again. The transmission consists of two parts: (i) the call-sign of the station GFP repeated in slow speed Morse for the first minute; (ii) the navigation signals which consist of a steady dash, interrupted at two points during the rotation of the beacon by the North signal (letter V followed by two dots) . . . — . . , and the East signal (letter B followed by two dots) — . . . . . To understand how these signals are used to obtain wireless bearings it is necessary to consider briefly the principles underlying the operation of the rotating beacon. Over a period of two or three years, experiments were carried out by the Radio Research Board with an experimental rotating beacon lent and erected by the Air Ministry for the purpose at Gosport on the South coast. These experiments have been described in detail in a previous report,* and from the results obtained it was concluded that this system was a reliable and useful aid to marine navigation, and that it would be desirable to subject it to a trial on the more extended scale in which the observations were taken by the personnel of the Mercantile Marine.

Accordingly, a more permanent type of beacon station was erected at Orfordness, and a general view of this station is shown in the photograph (Fig. 2). The rotating loop beacon comprises a closed loop or frame coil, which is supplied with current from a suitable radio transmitter. This loop, which is about 10 feet square, is mounted on a vertical spindle supported by a suitable stand, as shown in Fig. 3, and can be rotated about its vertical axis at a uniform speed. When such an installation is in operation, the rotating loop acts as the aerial for sending out wireless waves, but it has a characteristic feature that the strength of signal sent out in any direction varies during the rotation of the loop. The strength of signal actually obtained at a distant receiver varies according to the cosine of the angle between the plane of the coil and the direction of transmission, and it thus passes through a mininum value when the plane of the coil is perpendicular to the great circle through the transmitter and receiver. Hence the intensity of the signal heard in the telephones of the wireless receiver varies with time, passing through alternate maximum and minimum values with the rotation of the transmitting loop.

Method of taking bearings by stop-watch

In order to obtain bearings from such a rotating beacon it is necessary to determine the position of the loop at the occurrence of

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the signal minimum, and this is most easily carried out by a timing method. The transmitter is arranged so that a characteristic signal is emitted when the line of maximum radiation is in the N. direction—that is, when the plane of the loop is pointing E. and W. As already mentioned, the N. starting signal consists of the letter V followed by two dots, which is then followed by a long steady dash. If the operator at the receiver starts a stop-watch at the beginning of this dash and measures the interval which occurs until the received signal passes through its minimum, he can then determine his bearing from the transmitter with the aid of a knowledge of the time of rotation of the transmitter. The time of rotation of the Orfordness beacon is exactly 60 seconds, so that the coil moves at the rate of 6° per second. In any observing position, therefore, the bearing of the transmitter in degrees is obtained by simply multiplying by 6 the measured time in seconds from the N. signal to the minimum signal. Since the rotation of the transmitter is symmetrical about its axis, it is evident that two signal minima are experienced for each revolution of the beacon. Successive bearing observations may thus be made at half-minute intervals during the operation of the beacon. This enables a series of line bearings to be observed, from which, with a knowledge of the speed and course of the ship, the actual positions of the ship may be plotted, or, if the ship be stationary and observations can be carried out on two or more beacons, an accurate position fix can be obtained from the average of several bearings observed on each beacon. In certain cases where the receiver is situated directly N. or S. of the transmitter the characteristic signal would probably not be heard owing to the passage of the signal intensity through its minimum. To provide for such conditions, the transmitter is arranged to emit another characteristic signal after a rotation of 90°—that is, when the coil is passing an easterly direction. It is evident that bearings observed from this E. signal as a starting point are subject to a correction of 90°. To avoid the necessity of making the arithmetical calculation referred to above, special stop-watches are available, which are provided with a compass card dial, as shown in Fig. 4, from which the bearing either in degrees or in points of the compass may be read off directly. The watch is started with the second-hand in the zero position upon receipt of the N. signal from the beacon. As the signal passes through its minimum the bearing is read directly off the watch from the position of the seconds hand.

Use of an automatic recorder

In many circumstances it may be desirable to avoid the necessity of the continuous attention of an operator in obtaining bearing observations with a watch or chronometer, or it may be required

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FIG. 4.—*Special Stop-Watch used for taking bearings from a Rotating Beacon.*

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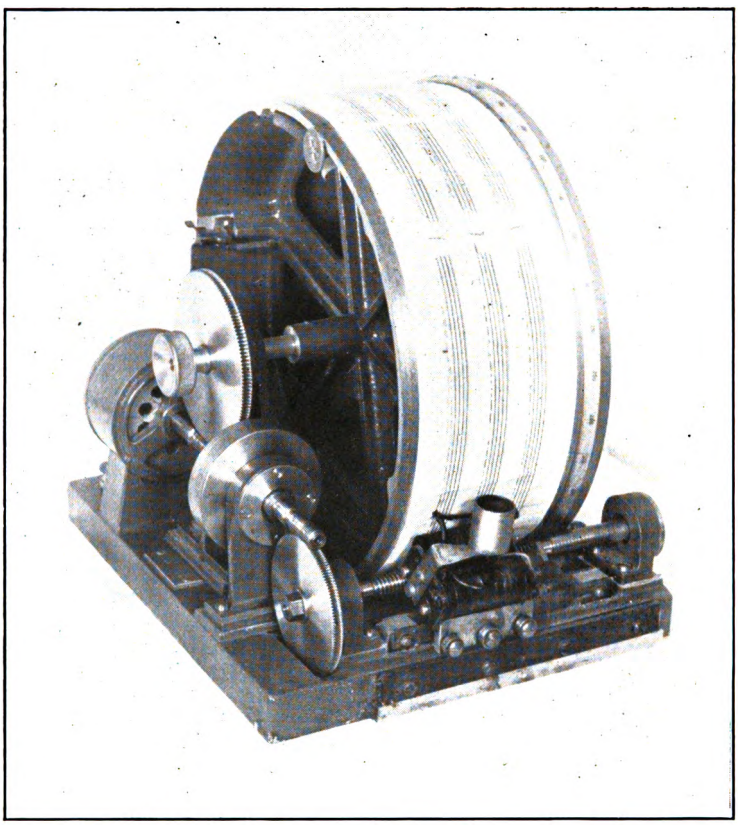

FIG. 5.—*Automatic Apparatus for recording bearings from a Rotating Beacon.*

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to obtain a permanent record of the bearings. For this purpose an automatic recording apparatus has been developed, a photograph of which is shown in Fig. 5. This recorder consists of a drum which is rotated at the same speed as the rotating beacon. The signals received from the beacon are caused to actuate a pen, which traces a spiral line on the rotating drum and records the signals as sent out from the beacon. During the five minutes transmission from a rotating beacon of the type now in operation at Orfordness, a repetition of the call-sign in Morse characters is recorded during the first minute. After this the record consists of a steady dash interrupted by the characteristic N. and E. point signals, and by gaps which occur in the trace at the positions of minimum signal intensity received from the beacon. A sample record taken of signals from the Orfordness beacon in this manner is illustrated in Fig. 6. When left in operation the apparatus will continue to record the received signals from the beacon over a period of half-an-hour or longer. At any time the received bearings may be read directly off the drum by setting the circular protractor scale provided so that the 0° reading coincides with the N. signal mark, and then noting the reading of the protractor scale at the mid-point of the gaps in the record which indicate the occurrence of minimum intensity of the received signal. When the record is removed from the drum the bearing corresponding to any minimum signal can be obtained directly from the paper with the aid of the graduations as shown in Fig. 6. From this sample record it can be seen that the observed bearings taken on a receiver in a fixed position are consistent to within 1°, which is about the overall accuracy of the rotating beacon transmitter.

When a second rotating beacon is in operation signals from this may be recorded on the same apparatus and on the same record if the transmissions are arranged to take place in the five-minute intervals between the consecutive transmissions sent out from Orfordness. In this way a record will be obtained of the bearings on two beacon stations so that the position of a ship may be plotted continuously over a period of half-an-hour.

In the experiments made up to the present time the recorder has been used at a distance of 93 miles overland from the beacon, and at this distance the accuracy of the bearings obtained is considered to be well within 1° under normal conditions of operation. When used on board ship under somewhat more favourable conditions, it is likely that the operation of such an instrument would be equally reliable up to ranges in excess of 100 miles. It is considered that the recorder could easily be developed along lines which would make it suitable for the installation in any part of the ship, preferably equipped with a receiver specially tuned in to receive signals from a rotating beacon.

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3. Results obtained from the Orfordness Beacon

Observations made in ships at sea

The previous investigations conducted by the Radio Research Board had provided information as to the reliability of the bearings obtained from such a rotating beacon when the observations were made by a skilled observer with the receiver either on land or in a ship at sea.

The first object of the erection of the Orfordness beacon was to ascertain what performance was obtainable when the observations were carried out by the personnel of the various classes of ship which would ultimately use the beacon signals as an aid to marine navigation. Accordingly, the Mercantile Marine Department of the Board of Trade invited the masters of ships and wireless operators to ascertain the bearing of their ships from the beacon by the aid of the received signals, especially when their ships were in known positions, and when bearings were also obtainable by visual or other means. Arrangements were made for the results to be recorded on a special form and sent to the Board of Trade for analysis. The results summarised and discussed below in this section were obtained in the above manner, and submitted by some 160 ships of all descriptions between July, 1929, and April, 1930.

Table 1 gives a summary of the results obtained by the Trinity House authorities with a receiver in a ship located in a series of fixed positions at or near various Buoys and Lightvessels in the south-western portion of the North Sea.

In this table it is seen that 16 out of the 21 observations recorded show no error in the observed bearing, and that the maximum in the remaining five cases is 1°. This refers to the case of bearings taken in a ship at anchor in such a position that its true bearing from the beacon is accurately known. Under such favourable conditions it is to be concluded that at ranges up to 45 miles the wireless bearings obtained from the beacon are accurate to within 1°. This conclusion is comparable with that obtained from observations on land in daylight at a range of over 90 miles, as mentioned above in connexion with the description of the use of an automatic recorder.

General results obtained at sea

In considering the results obtained by various ships at sea engaged upon their normal duties, attention must be directed to various factors which may affect the accuracy of the wireless bearings recorded. In the first place it appears from the reports that the majority of the ships were not provided with a 60 seconds stop-watch or chronograph suitable for making accurate bearing observations, and that the master or operator had to resort to the use of the seconds-hand of the ship's chronometer. Such an instrument is difficult to read to an accuracy of within half-a-second, and a calculation has to be made for the time of the north signal.

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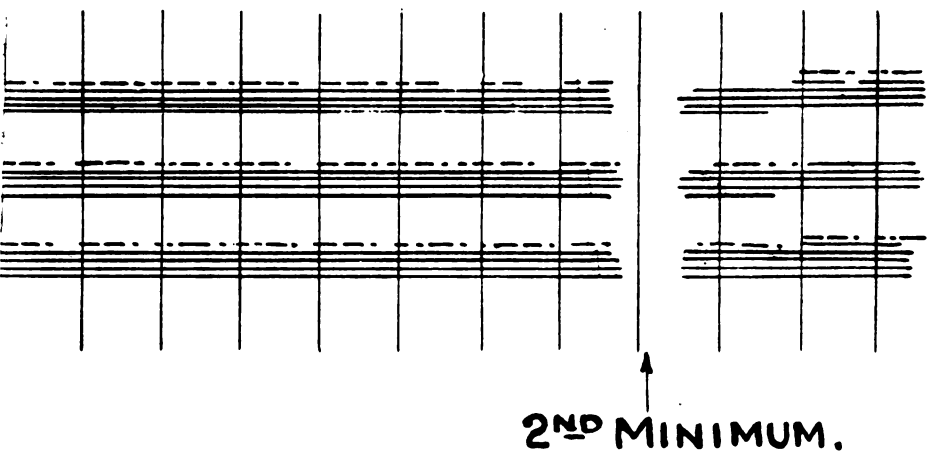

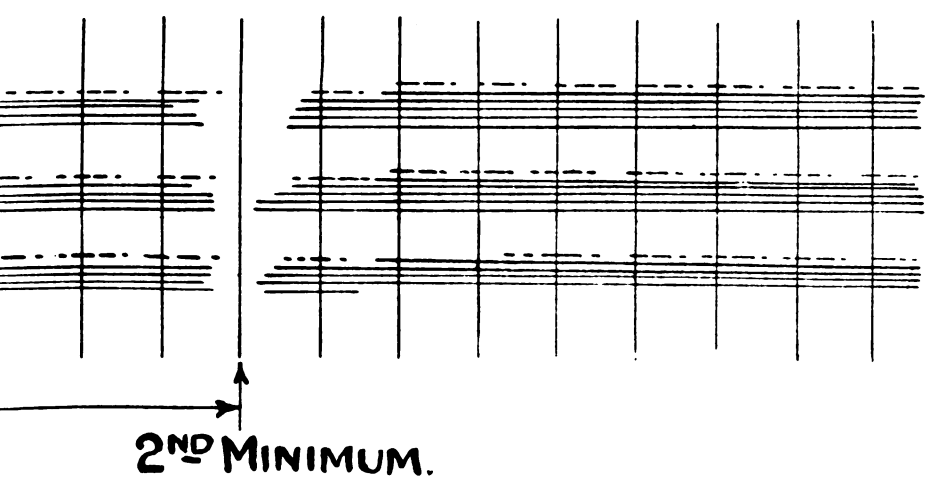

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TABLE 1.—*Results obtained by the Trinity House Authorities at various positions, 9th July to 9th August, 1929*

Position of Receiver	From Orfordness		Error in wireless bearing
	Distance	True bearing	
	Miles	Degrees	Degrees
——	4·0	326·0	0·0
Shipwash Light Vessel	8·5	115·0	− 1·0
Outer Gabbard Light Vessel	18·1	106·0	− 1·0
Galloper Watch Buoy	25·0	145·5	0·0
Barrow Deep Light Vessel	25·0	201·0	0·0
Black Deep Watch Vessel	29·0	195·0	0·0
N. Goodwin Light Vessel	45·0	180·0	0·0
Elbow Gas Buoy	43·0	181·0	− 0·5
Near Kentish Knock Light Vessel ..	26·4	168·0	0·0
At Kentish Knock Light Vessel	26·4	171·0	0·0
——	7·0	114·0	0·0
Outer Gabbard Light Vessel	18·1	106·0	− 1·0
S. Outer Gabbard Buoy	20·0	117·0	0·0
N. Galloper Buoy	21·5	135·0	0·0
Galloper Light Vessel	25·0	144·0	0·0
N. Goodwin Light Vessel	45·0	180·0	0·0
Off Margate Roads	40·0	183·0	0·0
5 miles S.S.W. of Kentish Knock Light Vessel	30·0	174·0	0·0
Longsand Head Gas Buoy	18·0	175·0	− 1·0
Galloper Light Vessel	25·0	144·0	0·0
Outer Gabbard Light Vessel	19·5	107·0	0·0

When it is realised that the beacon rotates at 6° per second, it is evident that this method is subject to a possible error of 3°. Further, in order to obtain the error in the observed wireless bearing, it is necessary to know the true bearing of the ship from the beacon: sometimes this can be obtained from visual observations, but in other cases resort must be had to dead reckoning, with a consequent decrease in the reliability of the bearing obtained. Any difference between the wireless and visual or calculated bearing must be ascribed to an error in the former, although previous experience has shown that this is by no means always the case. There is also the point that even when a stop-watch is employed it requires some little practice for an observer to attain the highest possible accuracy in reading the time of occurrence of the signal minimum after the N. signal.

In view of these facts, a general survey of all the reports received from ships indicates at once that the rotating beacon system is successful in providing the mariner with a new means of obtaining bearings to an accuracy which is usually quite adequate

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for his needs. At ranges up to 100 miles no difficulty has been experienced in obtaining bearing observations with the aid of the one-valve or two-valve type of receiver which is now fitted to the majority of ships.

Table 2 gives a summary of the records sent in by those ships which have taken more than 10 observations on the beacon in either one or more series.

TABLE 2.—*Summary of Bearing Observations on Orfordness Rotating Beacon, obtained by various ships between July, 1929, and April, 1930*

Ship	Number of observations	Distances from beacon Miles	Maximum error in wireless bearing Degrees	Percentage of bearings with error 2° or less
H.M.S. Adventure ..	433	13	8	69
H.M.S. Adventure ..	118	3	5	67
H.M.S. Argus	74	27 to 39	7	65
S.S. Bangalore	143	43 to 908	3·5	87
S.S. British Duke ..	24	19 to 49	4·5	58
S.S. British Scout ..	15	15 to 270	4	80
S.S. British Tommy ..	112	9 to 464	8	85
S.S. City of Palermo ..	52	8 to 110	3	79
S.S. Corpath	40	12 to 160	5·5	85
M.V. El Argentino ..	39	13 to 208	2	100
S.S. Godetia	14	8·5 to 49	4	71
S.S. Isle of Thanet ..	108	59 to 69	6	79
S.S. Jabiru	55	45 to 228	6	65
S.S. Panama Transport ..	78	55 to 291	4	94
H.M.S. Pangbourne ..	11	41 to 57	2	100
S.S. Perth	14	9 to 77	2	100
H.M.S. Renown	19	27	7	79
H.M.S. Rosemary ..	79	30 to 550	12	72
S.S. Suecia	58	9 to 196	3	97
Train Ferry No. 1 ..	21	13 to 78	4	95
Total	1,507	3 to 908	12	80

In most of the cases referred to in the table a stop-watch was employed, either of the usual type calibrated in fifths of a second, or of the special type already referred to and as illustrated in Fig. 4. Although, as will be seen from the table, the range of observation was over 900 miles in one case, the majority of the results were obtained within 100 miles of the beacon. From this mode of presenting the results it will be noted that, while in one

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case an error in bearing of 12° was recorded, in the majority of instances the maximum error was 4° or less. The variability in the results is undoubtedly due to the varying skill and experience of the observers, the fluctuating conditions as regards interference in the wireless reception, and the sum of the personal and watch errors. Altogether out of over 1,500 observations taken by observers on 19 ships, it is found that 80 per cent. of the wireless bearings are correct within 2°. This result compares very favourably with that obtained in the earlier experiments with the Gosport experimental beacon when, with skilled observers located in ships in accurately-known positions, it was found that 85 per cent. of all observed bearings were correct to within 2°. This limiting accuracy of 2° is comparable with that of the normal ship's wireless direction-finder, and is further considered to be adequate for most navigational purposes.

In the report* on the Gosport experiments it was stated that the reliable working range of that beacon for first-class bearings was about 50 miles with a normal type of ship's two-valve receiver, while slightly less reliable bearings were obtainable up to about 100 miles by day or night. The results obtained from the Orfordness beacon show that this is more efficient than the previous station, and the reliable working range of the beacon for first-class bearings may be stated to be at least 100 miles, while bearings of a somewhat lower order of accuracy are obtainable up to twice this range.

The reports contain several interesting records of "freak" reception of the beacon at exceptionally great distances. The most notable of these is that submitted by S.S. *Bangalore*. When in a position near Algiers, at a distance of 908 miles from Orfordness, and using a single valve detector only, the operator took a series of seven bearings, the mean of which was practically coincident with the true bearing. In a somewhat similar manner the S.S. *Sasebo*, when fishing off the West coast of Ireland, took bearings on the beacon at a distance of 491 miles. The S.S. *Hebrides* has also submitted bearings accurate to within 2° taken while the ship was at anchor at distances of 428 and 446 miles from the beacon.

Coastal deviation

When a series of experiments was carried out several years ago with a direction-finding station situated at Orford, about a mile inland from the present beacon, it was found that in the directions 225° to 240° the observed bearings were subject to a permanent error, which amounted to 2·5° on a wave-length of 1,000 metres. This error was attributed to a coastal deviation effect, since in the directions in which it occurred the waves travelled practically parallel to the coast line. It is natural, therefore, to inspect the results obtained on the beacon to ascertain whether any similar error is present in the bearings obtained from it. Definite evidence of the

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existence of such an error is available in the results under discussion, and also among those taken on land at a distance of some 93 miles from the beacon. The results show that in the approximate directions 22° and 240° from the beacon a permanent error of from 1° to 2° is experienced in the observed bearings over an arc of about 5° on either side of the above directions. Apart from this there is no evidence amongst all the results reviewed that any instrumental error is associated with the rotating beacon installation itself.

Night effects

The majority of the results recorded in Table 2 were taken in the daytime, and the summary thus represents the accuracy of bearings obtainable at the various ranges indicated. At distances up to 60 or 100 miles this accuracy will be maintained during the hours of darkness, but at greater ranges there is the possibility of the received signals being accompanied at night by variable phenomena which are commonly grouped under the term " night effects." In the first place, there is the possibility of fading of the received signal taking place and introducing false minima ; secondly, the signal minima become blurred or even non-existent in extreme cases, while on other occasions sharp signal minima may be obtained which, however, give bearings which are in error by variable amounts. It is known from previous experience in marine direction-finding that these effects are not usually encountered at distances of less than 70 miles, while the bearing errors do not become serious until the distance exceeds 100 miles, the transmission being oversea in both cases. The series of reports under review, and also a number of special observations carried out at fixed receiving stations on land, indicate that similar phenomena take place in the case of bearings obtained by signals from the rotating beacon.

The matter of the occurrence of errors in bearings during the hours of darkness is one to which the Radio Research Board has devoted much attention during the past few years, and it is probable that the future development of the rotating beacon system along lines which are already indicated from the results of this research will lead to a type of installation which is nearly, if not quite, free from the occurrence of these errors. In the case of the present installation it may be said that while at ranges in excess of 100 miles bearings taken at night should be viewed with caution, the most expedient way of obtaining a moderately accurate bearing under such conditions is to take a series of from 5 to 20 successive readings and calculate the average bearing from these. The means of such a series of bearings is not likely to be in error by more than 3° or 4° unless the night conditions are exceptional.

4. General Conclusions

From the discussion of all the results so far obtained, it is to be concluded that the inauguration of the rotating beacon service at Orfordness has achieved a large measure of success among those

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for whom it was intended, viz., the Mercantile Marine. The beacon has shown itself to be a most useful wireless aid to navigation, and is capable of providing bearings to ships at sea to an accuracy of the order of 2° when the observers have had some experience at the work. There is no evidence of any error greater than about 0·5° associated with the beacon installation itself; in directions parallel to the coast line at Orfordness, however, there is the possibility of a deviation of the bearing of the order of 2°. Otherwise, at moderate ranges the accuracy is limited entirely by the skill of the observer, and the type of watch or clock employed. On a conservative basis it may be stated that with the ordinary type of ship's two-valve receiver the reliable service range of the beacon is from 70 to 100 miles at all times, while the range of accurate bearings is increased to 250 miles in the daytime when conditions are favourable in regard to freedom from interference.

In the course of the experiments some observations have been made on ships fitted with only a crystal receiver, in which case the range of the beacon is found to be less than 10 miles. As, however, it is likely that in the very near future all ships will be fitted with a more modern type of valve receiver, this limitation of range will be removed.

As in the case of other systems of wireless direction-finding, experience has shown that the prevalence of fog or other bad weather conditions has no effect upon the observed bearings. With the rotating beacon system also, the observed bearings are independent of the yawing or rolling of the ship, and of the accuracy of its compass. The reliable daylight range given above will, therefore, be considered to hold under all weather conditions. During the hours of darkness, however, bearings taken at ranges exceeding about 100 miles are subject to certain difficulties and the possibility of error. A separate investigation of these effects at fixed stations has shown that the errors resulting from observations made under such conditions may sometimes be serious from the navigation point of view. Previous experience in the field of direction-finding, however, has indicated the means by which this limitation may be removed from the rotating beacon transmitter, and it is possible that with the continued development of the system future rotating beacons will have a reliable working range to the average ship's receiver of anything from 250 to 500 miles. Finally, in the use of the rotating beacon system for obtaining bearings at sea no special apparatus is necessary on board the ship other than the wireless receiver normally fitted, and a stop-watch (preferably of the special type described on page 4 and illustrated in Fig. 4).

In the Appendix which follows is given a selection of abstracts from the remarks which have accompanied the reports from many ships. These serve to illustrate the appreciation of the rotating beacon service by the mariner under practical sea-going conditions.

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APPENDIX

EXTRACTS FROM REPORTS FROM SHIPS

" I have taken bearings by this method, starting from the Nore right round to North of Yarmouth Roads ; also across to Antwerp and back, and have had very good results (distances up to 400 miles). The method is only guaranteed to about 80 miles and less than half that distance is all required for practical purposes. The signals are given out every alternate five minutes. So accurate to time is this that you have the correct time given for every five minutes."—*Captain, S.S. " Corland."*

The *M.V. El Argentino* obtained very good results from the Orfordness Rotating Beacon during very thick weather on the 8th November. The Captain reports that while he was unable to check these bearings by a known position of the ship, he found them to be exceedingly useful from a navigational point of view, and the bearings assisted him in maintaining a westward course during very unfavourable weather. In clearer weather when visual bearings were possible, the wireless bearings showed an error of less than 2° at all distances up to 200 miles.

" This system is very easy to work. I have tried personally, and was able to get satisfactory results. In my opinion it has great advantages over the usual direction-finder, in that the navigating staff can take bearings of the rotating beacon without any difficulty."—*Master, S.S. " British Scout."*

" These observations were taken by ordinary watch and found correct on all bearings. I strongly recommend that more of these beacons should be erected round the coast. I consider they are the finest aid to navigators yet established."—*Master, S.S. " Newton Abbot."*

" The signals from Orfordness rotating beacon are very strong and distinct at a distance up to 200 miles with a two-valve amplifier, but during the time broadcasting stations are operating these stations are disturbing too much to get bearings at longer distances than about 75 miles."—*Operator, S.S. " Suecia."*

" I had a bearing at a distance of 378 miles which was approximately 2° in error. I must add that these bearings were of great assistance to me in making my landfall, and will give me great confidence in future, should I be making Lundy Island in thick weather."—*Master, S.S. " Changuinola."*

" The Orfordness rotating wireless beacon has been thoroughly observed by the chief W/T Station at Norddeich and by the coastal W/T Station at Cuxhaven. The bearings obtained by these two stations may be said to be useful, taking into consideration the distance, which is considerable."—*German Ministry of National Defence.*

" In my opinion, this system has several advantages, most particularly not being dependent on the ship's compass when the vessel is rolling."—*Master, S.S. " British Tommy."*

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" Whilst coming down the North Sea last night we encountered a severe south-west gale, heavy head sea and also very heavy rain squalls. Many bearings were obtained from the new rotating wireless beacon at Orfordness, and these proved very satisfactory. As bearings from the beacon can be obtained as often as desired without any waiting for turn with excellent results, I consider this new method of obtaining wireless bearings better than previous methods. The bearings from this new station are free."—*Master, M.V. " British Reliance."*

" The above beacon can be heard with good signals off the North of Ireland about 500 miles distance, but is, of course, no use for navigation purposes on this route, Liverpool to Montreal. The zero is clear and can be observed with an accuracy of one second, which amounts to 6° in the bearing."—*Officer i/c S.S. " Laurentic."*

" A very valuable aid to navigation and more stations should be erected."—*Commander, S.S. " Laurel Branch."*

" On passage from Rouen to Tyne, the ship being then off Dungeness, a distance of 73 miles from the beacon, all the observations came within ½° of the correct result, the minimum intensity being at all times very clear. This installation has many advantages over the direction-finder, as it is less expensive and much more simple and accurate. The navigator has no worries regarding compass error or bad bearings due to the vessel rolling."—*Master, S.S. " Hetton."*

" The wireless beacon at Orfordness is one of the best aids to navigation on the East coast, and in thick or hazy weather I have had a number of bearings and found them most useful in different parts of the North Sea."—*Master, S.S. " Royston."*

" Transmission from beacon in all respects perfect: signals clear and strong."—*Master, S.S. " London."*

" I beg to place on record that I find the wireless beacon at Orfordness most useful now that it is working twenty-four hours per day . . . I hope it may be possible for you to make the present arrangements permanent, although I understand that they are so far only experimental."—*Master, S.S. " Suecia."*

After reporting an accurate observation from Orfordness while in the Locks at Avonmouth, distance 163 miles from the beacon :—

" I consider this very satisfactory indeed, and suggest that a similar station should be erected at Lundy Island."—*Master, S.S. " Changuinola."*

" It would seem that for short distances this type of beacon is quite satisfactory and able to give good results, even though the observing ship is yawing in bad weather, which would lessen the accuracy of the ship's d.f. to some extent."—*Master, S.S. " British Scout."*

" The following is the result which came in beautifully, and would have been very useful to me in fog, especially if bound south and making for the East Goodwin Light Ship: Observed bearing, 264°; true bearing, 264°; distance, 54 miles."—*Master, S.S. " Foreland."*

" Signals very clear and distinct. Bearings within 1° at 78 and 124 miles."—*Master and Operator, S.S. " Corminster."*

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" We think this method of obtaining W/T bearings to be the best known method for use on board ships (our ship is also fitted with a W/T. D.F. apparatus). Perhaps you can inform us if beacon Orfordness will be installed permanently, and if any other beacons of this kind will be installed in the future at other places."—*2nd Officer, S.S. " Dordrecht."*

" The Captain and other Officers were highly satisfied with the degree of accuracy obtained without the use of any special instruments, and expressed their desire for the erection of a similar beacon elsewhere which would thus enable them to obtain cross bearings."—*S.S. "Jebba."*

" I have repeatedly used this wireless beacon for obtaining bearings, and after a little practice those received with the aid of a stop-watch agreed very well with the bearings ascertained with a D.F. apparatus, and when Orfordness was in sight, with the compass bearings. With the D.F. receiving apparatus " Telefunken " system it is quite possible to receive and obtain bearings for a distance of more than 250 miles.

" In my opinion the Orfordness rotating wireless fog signal is particularly adapted for finding bearings, as the bearings obtained are true, and therefore you are independent of the wireless transmission and of any doubtful deviation of the compass there may be, as well as of errors in comparison between compass and D.F. apparatus."—*S.S. " Schwan."*

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DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH

RADIO RESEARCH

SPECIAL REPORT No. 11

THERMIONIC EMISSION

A SURVEY OF EXISTING KNOWLEDGE WITH PARTICULAR REFERENCE TO THE FILAMENTS OF RADIO VALVES

BY

W. S. STILES, PH.D.

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LONDON:
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PREFATORY NOTE

THE subject of thermionics has developed in a few years from a method of studying the electrical properties of matter to eing the basis of one of the world's widest commercial applications f physical science. The phenomena involved are fundamental, eing those on which are based theories of the constitution of natter as we at present perceive it. The literature on the subject s widespread, and much of it inevitably abstruse.

It was decided by the Radio Research Board on the advice of he Thermionics Committee, that for the furtherance of knowledge, or facilitating research and technical developments, a critical urvey of the literature of the subject would be of very real value.

The compilation of this survey was entrusted to W. S. Stiles, h.D., of the National Physical Laboratory. An endeavour has een made to include the most important papers up to December, 930. It will be observed that in the Bibliography which accom- anies the Survey a Decimal System of Notation has been employed, nd that gaps in the numbering have been left to enable the reader o insert later references approximately in their proper position.

E. H. RAYNER,
Chairman,
Thermionic Committee, Radio Research Board.

DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH,
16 OLD QUEEN STREET,
LONDON, S.W.1.
January, 1932.

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SECTION 0

GENERAL OUTLINE

PHENOMENA of a thermionic character had been observed many years before the nature of the effects was realised. The systematic study of the subject may be said to commence with RICHARDSON'S paper *On the Negative Radiation from Hot Platinum* (0.099) in 1901, and his comprehensive memoir, *The Electrical Conductivity imparted to a Vacuum by Hot Conductors* (0.100) in 1903. The attitude is here definitely adopted that the negative leak from a hot body to an auxiliary electrode *in vacuo* represents the simplest case of conductivity produced by matter at high temperatures. It is assumed that electrons, normally retained in the substance by a potential discontinuity at the surface, are able, with rise of temperature and consequent increase in the velocities of thermal agitation, to escape through the surface and constitute a current between the hot body and the auxiliary electrode. If a Maxwellian distribution of velocities of the electrons within the metal be assumed, and if there is a potential discontinuity Φ at the surface, then the number of electrons which pass out from unit area of the surface per second is given by

$$N\sqrt{\frac{kT}{2\pi m}}\exp\{-\Phi e/kT\}^*,$$

where k = gas constant for a single electron.
T = absolute temperature.
e = charge of an electron.
m = mass of an electron.
N = number of electrons per unit volume in the metal.

If all the electrons which are emitted are driven across to an auxiliary electrode, the anode, by means of a sufficiently intense electric field, the saturation current between the anode and the emitter so obtained will amount to

$$SN\sqrt{\frac{kT}{2\pi m}}\exp\{-\Phi e/kT\},$$

where S is the area of the emitting surface.

The experimental test and verification of the formula relating saturation current and temperature was carried out by RICHARDSON for platinum, carbon and sodium. He found that the conductivity

* To avoid confusion between the base of natural logarithms and the electronic charge, e^x is written exp (x).

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in vacuo between an electrically-heated filament of Pt or C and a cold auxiliary electrode was, in fact, essentially unipolar, a current being obtained only when the applied potential was such as to drive electrons away from the emitter. For sodium, secondary effects gave rise to a conductivity in the opposite direction, but this never exceeded a twentieth of the " negative " conductivity. The values of the constants in the formula

$$\left.\begin{array}{c}\text{Saturation current per unit area} \\ \text{in amps./cm.}^2\end{array}\right\} = AT^{\frac{1}{2}} \exp(-b/T)$$

derived as the final results, are as follows :—

	A	b
Pt	$1{\cdot}6 \times 10^7$	$4{\cdot}93 \times 10^4$
C	$1{\cdot}6 \times 10^{15}$	$7{\cdot}8 \times 10^4$
Na	$1{\cdot}6 \times 10^{12}$	$3{\cdot}16 \times 10^4$

RICHARDSON points out in the same paper that the product of b and the cube root of the atomic volume is nearly a constant for these three elements.

The hypothesis that the negative thermionic current is to be attributed to electrons ejected from within the substance of the emitter was almost immediately called in question by WILSON, *On the Discharge of Electricity from Hot Platinum* (0.110), who in an investigation of the effects of gases on the negative leak from platinum wires found that treatment of the Pt wire with nitric acid very considerably reduced the saturation current, the admission of a little hydrogen, however, bringing the current back to its former value. Although the introduction of nitrogen and water vapour gave the same current as obtained in a vacuum, provided ionisation by collision did not occur, for hydrogen a considerably larger current was obtained. WILSON drew the conclusion that the thermionic emission of platinum *in vacuo* is due to traces of hydrogen occluded on the surface of the platinum. WILSON further showed that a thermo-dynamic proof of the Richardson formula $i = AT^{\frac{1}{2}} \exp(-b/T)$ could be given, based on the analogy between the emission of negative ions by the hot body and the evaporation of molecules from a fluid. This proof did not require the electrons to come from within the metal, as assumed by RICHARDSON, and hence if the thermionic current were due to some surface action producing electrons, the formula relating measured saturation current and temperature would still have the form of the Richardson result.

While the emission from metals occupied the attention of English physicists, a number of investigations were being carried out by WEHNELT, *On the Discharge of Negative Ions by Glowing Metallic Oxides and Allied Phenomena* (0.119), on the discharge of negative ions by glowing metallic oxides. The results obtained by him are summarised in a paper of 1905 (0.120). Metallic oxides were found to give a negative emission similar to that from metals, which

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varied with temperature in the way required by Richardson's formula. WEHNELT did not give actual values of the emission constants, but for BaO and CaO RICHARDSON calculated them from Wehnelt's results :—

	A	b
BaO ..	$1 \cdot 15 \times 10^8$	$4 \cdot 49 \times 10^4$
CaO ..	$0 \cdot 72 \times 10^8$	$4 \cdot 28 \times 10^4$

A fairly extensive investigation of the current from a Nernst filament, including the variation with temperature of the negative leak in vacuum, was carried out by OWEN, *On the Discharge of Electricity from a Nernst Filament* (0.121), in 1904, and again, RICHARDSON calculated the emission constants which were not given explicitly by OWEN :—

Nernst filament. $A = 10^4$; $b = 4 \cdot 62 \times 10^4$.

The paper by RICHARDSON, *The Emission of Negative Electricity by Hot Bodies* (0.130), in 1904, summarises the state of the subject at the time and includes a table of emission constants (RICHARDSON, WILSON, OWEN, WEHNELT), with a critical discussion of these.

At the time of this review it was still questionable whether or not the emission from oxide-coated metals consisted merely of a secondary effect due to a lowering of the exit work for the metal by the oxide layer. DEININGER, *Emission of Negative Ions from Glowing Metals and from Glowing Calcium Oxide* (0.140), to throw light on this point, measured the emission of metal and carbon wires, and then coated the wires with calcium oxide. The emission from the coated wires was found to depend only on the oxide coating and not on the character of the underlying filament. HORTON, *On the Discharge of Negative Electricity from Hot Calcium and from Lime* (0.141), compared the emissions from platinum, calcium on platinum, and lime on platinum, in helium at 3·24 mm. pressure and found the following values for the constants in Richardson's formula :—

	A	b
Pt	$1 \cdot 6 \times 10^6$	$6 \cdot 1 \times 10^4$
Ca on Pt ..	$1 \cdot 7 \times 10^4$	$3 \cdot 64 \times 10^4$
CaO on Pt ..	$6 \cdot 4 \times 10^4$	$4 \cdot 79 \times 10^4$

MARTYN, *The Discharge of Electricity from Hot Bodies* (0.142), measured the increase in the emission from a hot platinum wire due to (*a*) the introduction of hydrogen, (*b*) coating the wire with lime, and found the two effects occurred independently.

JENTZSCH, *Electron Emission of Glowing Metal Oxides* (0.143), following up the work of DEININGER, obtained in an extensive investigation of the emission from oxides the following values for the constants in the Richardson formula :—

$$AT^{\frac{1}{2}} \exp(-b/T).$$

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He states that at the pressures worked with (about 1 mm.) the emission is independent of gas pressure.

Oxide of	A	b	ϕ in volts $= kb/e$.
Ba	141 $\times 10^{15}$	$41{\cdot}6 \times 10^3$	3·58
Sr	152 $\times 10^{15}$	$44{\cdot}9 \times 10^3$	3·87
Ca	129 $\times 10^{15}$	$40{\cdot}3 \times 10^3$	3·48
Mg	1·01 $\times 10^{10}$	$39{\cdot}5 \times 10^3$	3·40
			2·06
Be	0·31 $\times 10^{10}$	$23{\cdot}9 \times 10^3$	3·13
Y	5,590 $\times 10^{10}$	$36{\cdot}3 \times 10^3$	3·26
La	206 $\times 10^{10}$	$37{\cdot}9 \times 10^3$	3·21
Al	1·92 $\times 10^{10}$	$37{\cdot}3 \times 10^3$	3·15
Zr	1,970 $\times 10^{10}$	$36{\cdot}6 \times 10^3$	3·06
Th	10·5 $\times 10^{10}$	$35{\cdot}6 \times 10^3$	3·20
Ce	586 $\times 10^{10}$	$37{\cdot}1 \times 10^3$	3·02
Zn	0·0919 $\times 10^{10}$	$35{\cdot}1 \times 10^3$	3·02
Fe	1,060 $\times 10^{10}$	$56{\cdot}9 \times 10^3$	4·04
Ni	8,370 $\times 10^{10}$	$51{\cdot}2 \times 10^3$	4·41
Co	1,590 $\times 10^{10}$	$30{\cdot}2 \times 10^3$	2·60
Cd	0·112 $\times 10^{10}$	$30{\cdot}2 \times 10^3$	2·60
In	0·00105 $\times 10^{10}$	$22{\cdot}5 \times 10^3$	1·94

At this time both RICHARDSON, *The Ionisation produced by Hot Platinum in Different Gases* (0.150), and WILSON, *The Effect of Hydrogen on the Discharge of Electricity from Hot Platinum* (0.151), published accounts of further experiments on the effect of gases, notably hydrogen, on the negative emission from hot platinum. The general conclusion reached by these workers was that the large effects on the emission observed were to be attributed to surface actions, volume absorption being relatively unimportant. RICHARDSON regarded the surface action as a lowering of the exit work.

We have to notice at this period a discussion which seems to have been initiated by SODDY, *The Wehnelt Kathode in a High Vacuum* (0.160), in 1907, and which is of particular interest as it culminated in the elucidation of the effect of space charge on the maximum current obtainable between electrodes in a thermionic tube. SODDY was unable to reconcile the assumption of specific temperature emission from Wehnelt cathodes in high vacuum, with the diminution of emission current observed on improving the vacuum conditions. Both RICHARDSON (0.161) and WEHNELT (0.162) answered these criticisms, but SODDY (0.163) in a further communication described experiments with a Wehnelt cathode which tended to show that gaseous ionisation and not specific electron emission was responsible for the conductivity produced by the glowing oxide. LILIENFIELD (0.164), in a reconsideration of Soddy's experiments and other experiments of his own, made the suggestion that under good vacuum conditions the electrons in the space between the electrodes will tend to inhibit the passage of the thermionic current and

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an increased anode potential will be necessary to ensure saturation. When gas and therefore positive ions are present, however, the negative space charge is largely neutralised, and for given anode potential and temperature of the emitter the thermionic current is more than under vacuum conditions.

Despite difficulties in obtaining ideal conditions for thermionic emission, the use of the emission to measure temperature was considered by RICHARDSON, *The Application of the Ionisation from Hot Bodies to Thermo-metric Work at High Temperatures* (0.169), to be a promising possibility. The very consistent results obtained by DEININGER were used to illustrate how this might be done and the precision which should be attained, an error of not more than about 5° at 1,500° K. being indicated.

The theory put forward originally by RICHARDSON required that the electrons emitted by a hot body should have velocities distributed in accordance with the Maxwellian Law for the distribution of velocities among gaseous molecules. The first experiments made on the actual distribution of velocities among the emitted electrons were by RICHARDSON and BROWN, *The Kinetic Energy of the Negative Ions Emitted by Hot Bodies* (0.170), who measured the distribution for the components normal to the emitting surface. For platinum *in vacuo* the Maxwellian Law was found to hold good. Confirmation of the law for the velocity components tangential to the emitting surface was subsequently obtained by RICHARDSON (0.171). (See Section 4 for later papers dealing with the distribution of velocities among thermionic electrons).

It is of interest to note that RICHARDSON, *On Thermionics* (0.179), in a paper of 1909, first put forward the suggestion that the phenomena previously referred to as "the emission of electricity from hot bodies," "the leak from hot wires, etc.," should be termed "thermionics." The term took root, and is now the recognised expression in scientific literature. Richardson's paper is chiefly concerned with the calculation of the currents between emitting surfaces and collecting electrodes when the initial velocities are taken into account, the mutual repulsions of the electrons in the inter-space being, however, neglected. The cases solved are parallel planes, inclined planes, coaxial cylinders.

In 1908 attention was drawn to the temperature changes which may be expected to occur in bodies emitting electrons, owing to the energy required to expel the electrons through the potential discontinuity at the surface. WEHNELT and JENTZSCH, *Temperature Variations in the Electron Emission of Hot Bodies* (0.180), *Energy of Electron Emission of Hot Bodies* (0.181), made measurements of the cooling effect produced in this way for lime-coated platinum wires. The measurements were complicated by heating effects due to bombardment of the filament by positive ions. At the lower temperatures the expected cooling effect was observed, but its magnitude was about ten times that calculated from the exponential

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constant in the Richardson formula. SCHNEIDER, *Energy of the Electrons emitted by Hot CaO* (0.182), repeated the experiments of WEHNELT and JENTZSCH, and again found values for the cooling effect in excess of the calculated value.

Meanwhile RICHARDSON and COOKE, *The Heat Developed during the Absorption of Electrons by Platinum* (0.183), *The Heat Liberated during the Absorption of Electrons by Different Metals* (0.184), *The Absorption of Heat Produced by the Emission of Ions from Hot Bodies* (0.185, 0.186), were engaged in a series of investigations of thermal effects in metal filaments, produced by the absorption or emission of thermionic electrons. The measured cooling effect in tungsten was found to agree very satisfactorily with the computed value, and substantial confirmation of the theory was thereby obtained. Using Wehnelt cathodes, however, the anomalous results found by WEHNELT and JENTZSCH were confirmed. Later work on such cathodes by WEHNELT and LIEBREICH, *Energy of the Electron Emission of Hot Bodies* (0.187, 0.188), indicated that the emission current was subject to wide variations and these were accompanied by corresponding changes in the measured cooling effect. (See Section 3 for the further consideration of papers dealing with the thermal effects of thermionic emission.)

The proof that the negative ions emitted by hot bodies in high vacua are electrons, rests on determinations of the specific charge e/m. Measurements by THOMSON in 1899 had established the result for carbon filaments. In 1911, BESTELMEYER, *Path of the Electrons Emitted by Wehnelt Cathode in a Homogeneous Magnetic Field* (0.190), carried out a very careful investigation of the e/m value for the ions emitted by a Wehnelt cathode and obtained the result $1{\cdot}766 \times 10^7$ e.m. units, which represented probably the most accurate determination of the specific charge of an electron available at that time.

It had long been known that, in addition to emitting electrons at sufficiently high temperatures, hot bodies also gave a positive ion emission which was usually of a transitory character, decaying more rapidly with time the higher the temperature. By measurements of the ratio of charge to mass of the positive ions emitted by a large number of different substances (Pt, Pd, Ca, Ag, Ni, Os, Au, Fe, Ta, W, C, brass, steel, nichrome), RICHARDSON and HULBERT, *The Specific Charge of the Ions Emitted by Hot Bodies* (0.195), established that in every case the ratio had about the same value, showing that the ions were independent of the material of the emitter and must be due to some impurity. More accurate experiments subsequently performed by RICHARDSON, *The Positive Ions from Hot Metals* (0.196), showed that in the case of platinum, for the first hours of heating, potassium ions were emitted, followed in the final stages by an emission of sodium ions, with possibly a few ions of iron. The evidence on the nature of the positive ions was later summed up by RICHARDSON in his treatise (0.241). The

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conclusion reached was that the *initial* positive ion emission from hot metals consists of ions of the alkali metals and chiefly of potassium ions.

The view that the emission of electrons from sodium, potassium, calcium and the metallic oxides arose as a secondary effect of chemical reactions, was strongly urged by FREDENHAGEN, *Emission of Negative Electrons by Heated Metals* (0.200), *Behaviour of Wehnelt Cathode in Different Gases* (0.201), as a result of a number of investigations. PRING and PARKER, and PRING, *The Ionisation Produced by Carbon at High Temperatures* (0.202, 0.203), investigating the emission from carbon, found the saturation current decreased continuously as the pressure decreased, and with progressive purification of the carbon. They formed the conclusion that the high currents formerly observed for carbon by RICHARDSON were to be attributed to chemical reactions with the impurities in the carbon and with the residual gases. RICHARDSON, *The Origin of Thermal Ionisation from Carbon* (0.204), criticised Pring and Parker's conclusions.

The theory of a specific temperature emission of electrons was again thrown open to doubt by the work of these investigators. In a paper in 1913, however, in which LANGMUIR, *The Effect of Space Charge and Residual Gases on Thermionic Currents in High Vacuum* (0.210), records the results of investigations on the limitation of thermionic currents by space charge and the effects of traces of residual gases on the emission, the researches of FREDENHAGEN, PRING and PARKER are reconsidered in the light of the new knowledge available, and it is shown how space charge limitation and residual gas effects may vitiate the conclusions of these writers. LANGMUIR emphasised the precautions necessary to obtain reliable and significant results in thermionic measurements and his paper marks the beginning of an era of exact measurements in thermionics. LANGMUIR showed that if the anode potential V be kept constant, and the temperature raised above a certain value, the current obtained is independent of temperature and is given by the expression

$$f \,.\, V^{3/2} \,.\, e/m,$$

where f is a constant depending on the dimensions and arrangement of the electrodes.

A similar result was obtained by SCHOTTKY, *Action of Space Charge in Thermionic Currents in High Vacua* (0.211), at about the same time, and several years earlier CHILD, *Discharge from Hot CaO* (0.212), had derived the formula in connexion with positive ion emission. The modifications necessary when the initial velocities of the electrons are taken into account (these were neglected in the theory leading to the three halves power law) are dealt with by SCHOTTKY and others.

RICHARDSON, *The Emission of Electrons from Tungsten at High Temperatures: an Experimental Proof that the Electric Current in Metals is carried by Electrons* (0.213), described a further attempt

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to provide convincing proof of the pure temperature character of the negative emission from tungsten, in a paper in 1913. The tubes used were carefully prepared in accordance with the technique found necessary by LANGMUIR. The magnitude and persistency of the thermionic current obtained were such as to rule out the possibility of the electrons being produced by gas impacts on the surface, or by any chemical process involving consumption of the tungsten. K. K. SMITH, *Negative Thermionic Currents from Tungsten* (0.214), continued Richardson's work.

DUSHMAN, *Determination of e/m from Measurements of Thermionic Currents* (0.215), in the account of a series of experiments on the emission from tungsten under space charge limitation conditions, after pointing out the substantial confirmation of the law derived by CHILD, LANGMUIR and SCHOTTKY for the maximum current for a given anode voltage, remarks :—

" The perfect definiteness of the results obtained, which are independent of vacuum conditions after a sufficiently high vacuum has once been attained, the reproducibility of the observations even after allowing gas to enter the tube and then re-exhausting, and the quantitative agreement obtained, not only in the above experiments, but in all the experiments so far carried out in this laboratory, point to the existence of a pure electron emission *per se*, which is not a secondary effect due to chemical reactions, as assumed by a number of other investigators, and which is a function of the temperature only."

SCHOTTKY, *Influence of Structure Effects, in particular the Thomson Image Force, on the Electron Emission of Metals* (0.216), dealing with the effects to be expected owing to the constitution of the thermionic emission current as a stream of electrical particles and not a continuous flow, reaches the interesting conclusion that the maximum current obtainable from a surface emitting electrons will be dependent on the electric field at the surface. This " Schottky Effect " was detected experimentally by SCHOTTKY and his theory confirmed. (See Section 5 for further papers dealing with the effects on the emission of an applied electric field at the surface.)

In 1914, LANGMUIR, *The Electron Emission from Tungsten Filaments containing Thorium* (Title alone published) (0.220), gave a description before the American Physical Society of an effect on the emission from a tungsten filament of a small quantity of thorium, present originally in the tungsten filament as thoria. The emission was in some cases increased by many thousand times.

A number of brief references to the effect were made by LANGMUIR, *Langmuir* (0.221), *The Relation between Constant Potentials and Electro-chemical Action* (0.222), *The Constitution and Fundamental Properties of Solids and Liquids* (0.223), in the period 1914–1923, but the detailed discussion of his experiments was deferred until 1923. (See Section 7 for further papers on the emission from thoriated filaments and other thin film emitters.)

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An investigation by SCHLICHTER, *Spontaneous Electron Emission from Glowing Metals and the Thermionic Element* (0.230), of the " spontaneous " current between a hot emitter and a cold receiving electrode practically surrounding the emitter, as compared with the saturation current obtained by applying a potential difference across the electrodes, is of interest. For platinum the two currents were found to be nearly the same at all temperatures worked with, and the absence of any considerable reflection of electrons is indicated.

A valuable review of the investigations on thermionics in the years 1905 to 1914 was published by SCHOTTKY, *Review of Thermal Emission of Electrons* (0.240) in 1915. In the following year RICHARDSON'S treatise, *The Emission of Electricity from Hot Bodies*, 1st Edition (0.241), appeared. A further comprehensive account of the subject was given by RICHARDSON, *Thermionic Electrodes* (0.242) in 1917.

The values of the constants A and b in the Richardson equation $AT^{\frac{1}{2}} \exp(-b/T)$, as measured by different investigators in high vacua in the presence of gases, after various treatments of the filament, etc., were shown by RICHARDSON, *The Influence of Gases on the Emission of Electrons and Ions from Hot Metals* (0.250), to conform to a simple empirical relation, $\log A = \alpha b + \beta$, where α and β represent constants characteristic of the emitting substance (platinum or tungsten).

A theory of the effect of gases on emission is put forward by RICHARDSON, which leads to the empirical equation just given.

It was pointed out by CASE, *New Strontium and Barium Photo-electric Cells* (0.270, 0.271), that the saturation current in an audion tube containing a filament coated with alkaline earth oxides is increased by illumination of the filament. Case attributes the effect to photo-electric emission from or by alkaline earth metals formed on the filament by reduction of the oxides. (See Section 6 for further papers dealing with the effect of light on thermionic emission.)

The theoretical relation between contact difference of potential and thermionic emission had been derived in the early development of the theory which led to the formula

$$V = (kT/e) \log i_2/i_1$$

where V is the contact potential difference between two surfaces at absolute temperature T and i_1 and i_2 are the thermionic saturation currents per unit area at the same temperature. An experimental proof of the result was given by RICHARDSON and ROBERTSON, *Contact Difference of Potential and Thermionic Emission* (0.280), who determined the change in the contact difference of potential between a thoriated tungsten filament and the cold anode, on activating the filament. The shift in the current-anode voltage characteristic gives immediately the change in the contact difference

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of potential, assuming the anode to be unaffected by the activation The theoretical formula was satisfactorily confirmed.

The effect of intense electric fields on the thermionic saturation current and other phenomena involving the potential conditions and microstructure of the emitting surface were discussed by SCHOTTKY, *Cold and Hot Electron Discharges* (0.290), in non-mathematical terms. The paper forms a fairly complete account of the author's views on the image force and its manifestations.

The introduction of cæsium vapour into a thermionic tube was found by LANGMUIR and KINGDON, *Thermionic Effects caused by Alkali Vapours in Vacuum Tubes* (0.300), to modify the electron emission in a characteristic fashion. (Papers on this subject are dealt with in Section 7.)

A decrease in the emission current from a tungsten filament when run at temperatures approaching the melting point was observed by JENKINS, *On the Emission of Positive Ions from Hot Tungsten* (0.310). This is related by JENKINS to the evolution of positive ions by the tungsten. This positive ion emission is of an entirely different character from that previously observed at lower temperatures, and in the presence of gases.

A general account of thermionic phenomena is contained in a monograph by BLOCH, *Thermionic Phenomena* (0.320), published in 1923.

Certain substances containing iron and alkali metal were investigated by KUNSMAN, *The Thermionic Emission from Substances containing Iron and Alkali Metal* (0.340), *A New Source of Positive Ions* (0.341), who found them to be constant sources of both positive and negative emission. The positive emission was identified as singly-charged ions of the alkali metals. Both emissions followed Richardson's equation, $AT^{\frac{1}{2}} \exp\{-\phi e/kT\}$. Corresponding values of ϕ for the positive and negative emissions in a given case were 3·41 and 4·00 volts respectively. (See 0.195, 0.196.)

MITRA, *On the Emission of Positive Electricity from Hot Tungsten in Mullard Radio Valves* (0.360), carried out a similar investigation to that of JENKINS on the emission of positive ions from tungsten at high temperatures, using commercial radio valves.

BREWER, *Factors Influencing Thermionic Emission* (0.390), *The Relation between Temperature and Work Function in Thermionic Emission* (0.391), observed the emission from gold and iron at atmospheric pressure of various gases, and found the positive and negative currents satisfied Richardson's equation, there being a relationship between the constants of the negative and positive emissions under different conditions.

An important treatise on thermionic emission appeared in 1928 as a volume of the *Wien-Harms Handbuch der Experimental Physik* (0.400). The work is divided into three sections, "Physics of Thermionic Emitters," by SCHOTTKY and ROTHE, "Preparation of Thermionic Electrodes," by SIMON, and "Technical Electron

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Tubes and their Application," by ROTHE. The first section includes a very comprehensive treatment of the thermo-dynamic theory of electron emission with detailed comparison with the experimental results. The effects of surface layers of other substances and the atomic and image fields at the surface are fully considered.

It was observed by JONES and DURAN, *Electron Emission at the Surface of Platinum through which Hydrogen is Passing* (0.405), that if hydrogen is passed through the walls of a hot platinum tube the emission of electrons from the exit surface far exceeds the Richardson current appropriate to platinum at the temperature concerned. The effect was complicated by surface actions similar to those orignally observed by WILSON (0.110).

Positive ion emission from tungsten, molybdenum, tantalum and rhodium was detected by WAHLIN, *The Emission of Positive Ions from Metals* (0.410), when the temperature of the metal was sufficently high for vaporisation to become appreciable. (Compare JENKINS (0.310) and MITRA (0.360).) An initial transitory emission of alkali metal ions occurred. SMITH, *The Emission of Positive Ions from Tungsten and Molybdenum* (0.420), working on the same effect, used a mass spectrograph and showed that at moderate temperatures (1,600–2,000° K.) tungsten and molybdenum emit sodium and potassum ions. Above 2,000° K. aluminium ions appear, and finally at about 2,500° (W) or 2,300° (Mo) ions of the heated metal itself are emitted. The work functions for the metal ions were determined by temperature variation measurements, as $\phi = 6{\cdot}55$ volts (W), $\phi = 6{\cdot}09$ volts (Mo). These values differ widely from those found by assuming the sum of the work functions of the electron and positive ion to equal the ionisation energy plus the heat of evaporation of the neutral atom. It is suggested that the ions are formed as a by-product of an irreversible recrystallisation of the metal.

A brief sketch of the development of the theory and experimental measurement of thermionic emission from metal surfaces is contained in a Dutch paper by ZWIKKER (0.425).

The thermionic properties of tungsten in iodine vapour were examined by KALANDYK, *Electric Emission of Incandescent Tungsten in an Atmosphere of Iodine* (0.430), who found a pronounced augmentation of the negative emission, decreasing with rise in filament temperature, and increasing with the pressure of iodine. KALANDYK attributes the extra emission to negative ions of iodine.

A critical review of the whole subject of thermionic emission is contained in a monograph by DUSHMAN, *Thermionic Emission* (0.450), published in October, 1930.

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SECTION 1

THEORY OF THE TEMPERATURE EMISSION OF ELECTRONS

THE first theoretical derivation of the expected relation between the number of electrons emitted by a hot body and its temperature, was given by RICHARDSON, *On the Electrical Conductivity Imparted to a Vacuum by Hot Conductors* (1.100), on the basis of the classical electronic theory of metallic conduction. The interior of the metal is treated as a region of constant potential, within which the electrons move about in temperature motion precisely as the molecules of a perfect gas at the same temperature. An electron reaching the surface of the metal can escape only if its velocity normal to the surface is sufficient to carry it through a certain potential discontinuity at the surface Φ. The distribution of velocities being Maxwellian, the number escaping in unt time from unit area of the surface is given by

$$N\,(kT/2\pi m)^{\frac{1}{2}} \exp\,(-e\,\Phi/kT)$$

where N = number of electrons per unit volume in the metal.
T = absolute temperature.
k = gas constant for a single molecule ($1\cdot37\times10^{-16}$ c.g.s. units).
e = charge on an electron ($4\cdot77\times10^{-10}$ c.g.s. e.s. units).
Φ = potential discontinuity at surface in c.g.s. e.s. units.

If a field is applied to collect all the electrons ejected, the thermionic current obtained is then

$$Se\,N(kT/2\pi m)^{\frac{1}{2}} \exp\,(-e\,\Phi/kT)$$

where S is the area of the emitting surface. This is Richardson's original derivation of the thermionic emission equation.

It is also pointed out in the same paper that the emitting body will lose energy owing to the emission, of amount $e\,\Phi + 2kT$ per electron. This results in a cooling of the metal (the cooling effect). (The experimental determinations of the cooling effect are dealt with in Section 3.)

It is implied in Richardson's analysis, although not pointed out explicitly, that the electrons after leaving the metal will have a Maxwellian distribution of velocities corresponding to the temperature of the emitter. (See Section 4 on the distribution of velocities among the emitted electrons.)

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An expression for the thermionic emission current, of the same form as Richardson's, was derived by WILSON, *On the Discharge of Electricity from Hot Platinum* (1.101), treating the ejection of electrons as a process analogous to the evaporation of a liquid, and applying the Clapeyron equation. WILSON finds that the electron current equals $AT^{\frac{1}{2}} \exp(-Q/2T)$, where A is a constant independent of temperature and Q is the internal work done in evaporating one gramme molecular weight of the electrons (expressed in small calories).

Use had previously been made by WILSON, *On the Electrical Conductivity of Air and Salt Vapours* (1.102), of thermodynamical methods, in deriving a similar formula to the above, for the current in an allied problem.

A further discussion of the theory of thermal emission of electrons was given by DEBYE, *Theory of Electrons in Metals* (1.110), who drew attention to the relation connecting the Volta difference of potential between different metals, their thermionic emission constants and their Peltier effect. The substance of Debye's paper is as follows. For two different emitters in temperature equilibrium in an evacuated enclosure, the potentials ϕ_1 and ϕ_2 just outside the two bodies will in general be different, and the concentrations n_1 and n_2 in the electron atmosphere at these places will be given by

$$n_2/n_1 = \exp\{e(\phi_2 - \phi_1)/k\mathrm{T}\}.$$

Also according to classical electron theory the concentration of electrons inside the two bodies, N_1 and N_2 will be related to those just outside, by the equations:

$$n_1 = \mathrm{N}_1 \exp\{-e\Phi_1/k\mathrm{T}\}$$
$$n_2 = \mathrm{N}_2 \exp\{-e\Phi_2/k\mathrm{T}\},$$

where Φ_1 and Φ_2 are the potential discontinuities at the respective surfaces.

Thus

$$\phi_2 - \phi_1 = k\mathrm{T} \log_e \frac{\mathrm{N}_2}{\mathrm{N}_1} - (\Phi_2 - \Phi_1).$$

The difference of potential $\phi_2 - \phi_1$ is identified as the Volta difference of potential between the metals, V. If these be assumed in contact another relation between N_2 and N_1 may be assumed, namely

$$\mathrm{P} = k\mathrm{T} \log_e \frac{\mathrm{N}_2}{\mathrm{N}_1},$$

where P is the Peltier coefficient. Thus—

$$\mathrm{V} = \mathrm{P} - (\Phi_2 - \Phi_1)$$

DEBYE also discusses the nature of the forces giving rise to the exit work $e\Phi$ of the electrons. A well-known result of electrostatic

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theory states that a point charge placed near a plane-conducting surface induces on the latter a charge distribution of opposite sign, such that the force between the point charge and the surface equals the force between the original point charge and an equal and opposite charge placed at the point where the optical image would be formed by reflection in the surface. This force is known as the Thomson image force. DEBYE points out that the Thomson image force between an electron and the plane-conducting surface from which the electron originated, may be regarded as accounting for the whole of the exit work, provided the image force is assumed to act only up to a certain critical distance x_0 (depending on the substance of the emitter) from the surface. The exit work is then given by $e^2/4x_0$. This conception is used to show that the Volta difference of potential is for all practical cases independent of the form of the conductors.

The question of the relationship between thermionic and thermoelectric effects was taken up by RICHARDSON, *On the Electron Theory of Contact Electro-motive Force and Thermo-electricity* (1.111). *Some Applications of the Electron Theory of Matter* (1.112), in papers published in 1912. Debye's result is confirmed and a new relation derived connecting the Thomson specific heat of electricity σ, the work done by an electron in escaping from the metal w, and the ratio of the specific heats of electron gas at constant volume and constant pressure, γ. Richardson's formula is as follows

$$\frac{\partial w}{\partial \mathrm{T}} = \frac{k}{\gamma - 1} - e\,\sigma.$$

It is obtained by consideration of a system consisting of two emitters of the same metal but at different temperatures, each in equilibrium with electron gas of appropriate pressure and temperature, and joined together by a thin wire which permits of thermal readjustment only infinitely slowly. RICHARDSON also discusses the possibility of thermionic emission being a type of auto-photo-electric emission brought about by the temperature radiation of the hot body. By means of a thermodynamic argument it is shown that whatever the mode of origin of the electrons (thermionic, photo-electric, etc.) the number per unit volume in equilibrium with an emitter at temperature T will be given by

$$\mathrm{A} \exp \left\{ \int_0^{\mathrm{T}} \frac{w}{k\mathrm{T}^2}\, d\mathrm{T} \right\}.$$

where w is the latent heat of evaporation of the electrons.

The equation $\frac{\partial w}{\partial \mathrm{T}} = \frac{k}{\gamma - 1} - e\sigma$ provides a method for determining the variation of w with temperature. RICHARDSON points out that σ is small compared with $\frac{k}{\gamma - 1}$ (γ put equal to 5/3) so that

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$w = w_0 + 3/2\,kT$. This leads to an emission current equation $i = AT^2 \exp(-w_0/kT)$ where A and w_0 are independent of temperature. This suffices to represent the experimental data to as good an approximation as does the original Richardson expression

$$i = AT^{\frac{1}{2}} \exp(-w_0/kT).$$

In deducing the expected emission current it is customary to multiply the number of electrons per unit volume in equilibrium with the hot body by

$$\left(\frac{kT}{2\pi m}\right)^{\frac{1}{2}}.$$

RICHARDSON here points out that this assumes the absence of electron reflection and that in practice a smaller current will be obtained than that derived theoretically, if any appreciable reflection takes place.

Criticisms of Richardson's analysis were made by BOHR, *Note on the Electron Theory of Thermo-electric Phenomena* (1.113), and H. A. WILSON, *Note on the Application of Thermodynamics to the Discharge of Electricity by Hot Bodies* (1.114). Bohr's criticisms consist in pointing out that the Thomson and Peltier effects, as measured, are produced by a current flowing uniformly under an applied potential gradient and are not necessarily identical with the corresponding effects which are produced by the virtual displacement of electrons under equilibrium conditions as assumed in the theoretical deduction. RICHARDSON, *On the Electron Theory of Thermo-electric and Thermionic Effects* (1.115), in answer to the criticisms, slightly modifies the meanings to be attached to w and σ.

W. WILSON, *Attempt to Apply the Quantum Hypothesis to the Electrical Discharge of Hot Bodies* (1.120), in a paper of 1913, adopts the standpoint that the emission may be regarded as a type of photo-electric effect brought about by the temperature radiation of the hot body. By considering the equilibrium in the metal, between emitted and adsorbed radiation and electrons, treating atoms as Planck oscillators and applying the quantum theory notions of discontinuous energy exchange, the expression

$$T\left\{1 + 2\frac{k}{w}T + \frac{2k^2}{w^2}T^2\right\} \exp(-w/kT)$$

is derived for the thermionic emission current, where w is the minimum energy possessed by an electron in order for it to escape from the body. This expression is shown to fit the experimental data equally well with the Richardson equation

$$i = AT^{\frac{1}{2}} \exp(-w/kT).$$

The application of quantum theory to the thermionic effect is dealt with in another way by RICHARDSON, *The Distribution of the*

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Molecules of a Gas in a Field of Force with Applications to the Theory of Electrons (1,121), in a paper of 1914. For the thermionic saturation current the following expression is derived—

$$0{\cdot}472\,\frac{2\sqrt{2\pi}}{9}\left(\frac{3e}{5}\right)^{3/2}\frac{Mk^2e}{Nh^3}\,T^2\exp\left\{-\frac{w_2-w_1}{kT}\right\},$$

where M = molecular weight of electron gas and N is Avogadro's constant. The interesting feature of this result is that, writing it in the form $AT^2\exp\{-b/T\}$, A and b are independent of temperature and A is the same for all metals, i.e. a universal constant. The value of this constant is given by RICHARDSON; $A = 1.47 \times 10^{-10}$ in e.s. c.g.s. units.

The suggestion of DEBYE, that the exit work of an electron may be due to the attraction between it and its electrical image in the emitting conductor, was further investigated by SCHOTTKY, *Influence of Structure Effects, in particular the Thomson Image Force, on Electron Emission of Metals* (1.130). SCHOTTKY points out that, owing to the image force, electrons leaving the metal will not all escape, some being pulled back into the conductor. The image force falls off rapidly with the distance out from the conductor, so that electrons which reach a certain plane at about 10^{-4} cm. from the surface will practically all escape. The application of an electric field at the surface to draw the electrons away to a collector electrode will not only draw away those passing beyond the plane at 10^{-4} cm., but will also enable some to escape which the image force would have recaptured in the absence of the external field. The effect is to shift the critical plane nearer to the metal surface. The net result is a slight diminution of the exit work, so that in place of the emission formula

$$i = Af(T)\exp(-b/T)$$

for zero external field, the modified expression

$$Af(T)\exp\left\{-\frac{(b-\Delta b)}{T}\right\}$$

must be substituted where

$$\Delta b = \frac{e^{3/2}}{k}\sqrt{\frac{dV}{dx}} = 4{\cdot}39\sqrt{\frac{dV}{dx}}$$

and $\frac{dV}{dx}$ is the external electric force at the surface in volts/cm. SCHOTTKY describes preliminary measurements with tungsten which confirm the theoretical variation of emission current with the electric force at the emitter surface.

The formula

$$\frac{\partial w}{\partial T} = \frac{k}{\gamma - 1} - e\sigma$$

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derived in 1912 by RICHARDSON, is criticised by SCHOTTKY, *Temperature Dependence of the Potential Discontinuity Metal Vacuum* (1.140), on the ground that the heat absorbed when an electron is evaporated from a metal and simultaneously another electron flows along inside the emitter to take its place, is set equal to the heat absorbed when an electron is evaporated from an insulated emitter. The validity of this assumption is called in question by SCHOTTKY, who derives the modified equation

$$\frac{\partial w}{\partial T} = \frac{k}{2} - e\sigma - e\frac{dV^1}{dT} + \frac{dA}{dT},$$

where

$$\frac{dV^1}{dT} \text{ is the limit of } \frac{V_2 - V_1}{T_2 - T_1} \text{ as } T_2 \to T_1 = T,$$

and $V_2 - V_1$ is the Volta potential difference between similar emitters at temperatures T_2 and T_1 respectively. $\frac{dA}{dT}$ is a term which depends on the conditions of the electrons in the interior of the metal and equals k, assuming them to behave as the molecules of an ideal gas.

SCHOTTKY adopts the point of view that in the metal both free and bound electrons are present, and that the difference between the two latent heats is accounted for in part by the heat of dissociation of the bound electrons.

RICHARDSON, *The Complete Photo-electric Emission* (1.150), returned, in an address to the British Association in 1915, to the question of the possibility of the total photo-electric emission being identical with thermionic emission. He carried out a calculation of the probable magnitude of the saturation current from platinum at 2,000° K., due to total photo-electric emission throughout the interior of the metal, and obtained the figure $i = 2 \times 10^{-11}$ amps./cm.^{2} The observed thermionic emission for platinum at this temperature is of the order of 6×10^{-4} amps./cm.^{2} (LANGMUIR). RICHARDSON concludes that total photo-electric emission can give rise only to an insignificant portion of the total emission in thermionic measurements.

A critical discussion of the theoretical aspects of thermionics is contained in the review on the subject published by SCHOTTKY, *Review of Thermal Emission of Electrons* (1.160), in 1915. In the following year RICHARDSON'S treatise, *The Emission of Electricity from Hot Bodies*, 1st Edition (1.161), appeared containing a unified account of most of the earlier work on thermionics. Another general discussion on the theory of thermionic phenomena, due to LANGMUIR, *The Relation between Contact Potentials and Electrochemical Action* (1.162), also appeared at this time. LANGMUIR is particularly concerned with the nature of the work function w in the emission equation, and the assumptions of a double layer at the surface and of image forces are examined in detail.

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A modification of Richardson's original equation for the thermionic current was made at this time by HAUER, *Electrical Conductivity of Metals* (1.169), who, by applying the Planck law for the distribution of energy of the electrons to the dissociation equilibrium of ions, electrons and atoms, showed that N should vary as $T^{\frac{1}{2}}$. The emission equation under these conditions is of the form $i = AT \exp(-b/T)$, where A and b do not vary much with T.

The thermodynamic arguments of RICHARDSON and others depend on the assumption that an emitter in equilibrium with an atmosphere of electrons is precisely analogous to an ordinary solid body in equilibrium with its vapour. A comprehensive inquiry into the validity of this analogy was carried out in 1918 by LAUE, *Thermionic Electrons* (1.170). LAUE considers a number of emitters at the same temperature in equilibrium with an electron atmosphere of which they form the boundary, and employs the electrostatic and statistical relations

$$\nabla^2\phi = -\rho \text{ and } \rho = \rho_0 \exp\{-e\phi/kT\},$$

where ρ = density of the electronic charge } macroscopic
and ϕ = electrostatic potential } magnitudes,
to obtain the equation

$$\nabla^2\phi = -\rho_0 \exp(-\alpha\phi)\ ;\ \alpha = e/kT,$$

which holds good at each point in the electron atmosphere filling the otherwise vacuous space surrounding the emitters. (*Note.*—LAUE uses Lorentz notation.) A solution of this equation represents a possible equilibrium distribution of electrons. The density of the electron atmosphere at the surfaces of the emitters is regarded as determined solely by the temperature and the nature of the emitting surface. For a region totally enclosed by emitting surfaces the electron density at every point inside is then uniquely determined. The interesting result then follows that for an enclosure of given size the electron density at all parts in it except those in the immediate neighbourhood of the enclosing emitting surfaces, is independent of the nature of these surfaces and is proportional to the absolute temperature, provided the latter exceeds a certain value, which, however, will depend on the thermionic properties of the walls. This state of affairs is quite different from that contemplated in the thermo-dynamic considerations of Richardson and others.

LAUE proceeds to determine the equilibrium distributions in certain simple geometric arrangements of emitting surfaces. He finds for the variation of density between two infinite plane-emitting surfaces situated at $x = x_0$ and $x = x_0 - 2\pi/\alpha K$

$$\rho(x) = \frac{\alpha K^2}{2\sin^2\left\{\frac{\alpha K}{2}(x - x_0)\right\}}$$

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which degenerates to

$$\rho(x) = \frac{2}{\alpha(x - x_0)^2},$$

when the second plane is removed to infinity ($K = o$). The normal pressure on the plane at $x = x_0$ is the sum of the electric traction due to the electrostatic field and the pressure of the electrons impinging against the surface. LAUE shows that these two balance out, and hence no work is done in displacing a plane, emitting surface. If, however, the surface is slightly curved, having two principal radii of curvature, R_1 and R_2 there is a net normal pressure or traction of the surface equal to

$$\frac{2}{\alpha}\sqrt{\frac{2\rho}{\alpha}}\left(\frac{1}{R_1} + \frac{1}{R_2}\right).$$

The accompanying figure (1.170*a*) indicates the direction of the force in the two cases which arise, namely

$$\frac{1}{R_1} + \frac{1}{R_2} > o \text{ and } \frac{1}{R_1} + \frac{1}{R_2} < o.$$

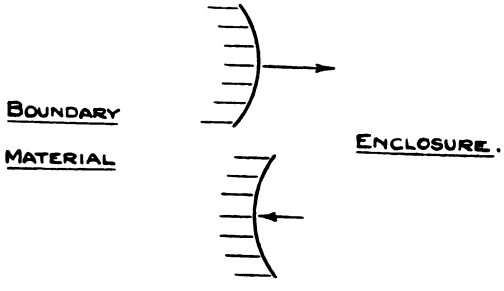

FIG. 1·170*a*.

Thus, in the expansion of any finite enclosure, for which it cannot be true that at all points $\frac{1}{R_1} + \frac{1}{R_2} = o$, the electron atmosphere inside will do work. This permits of the application of thermodynamics to such a system, and LAUE, by considering the change of entropy of the electron atmosphere between emitting electrodes, is able to derive the equation for the electron density at the boundary of the emitting surface,

$$\rho_o = \left(\frac{Ne}{2R}\right)^3 C^2 T^{3/2} \exp(\mu/RT),$$

where $N = 6{\cdot}2 \times 10^{-23}$ (Avogadro's Number)
μ = thermodynamic potential of electrons in the equilibrium system electrode-atmosphere, per mol. of electrons.
$R = Nk$
C = a constant.

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Remembering that $\rho_g = ne$ and assuming Maxwellian distribution of velocities among the electrons of the electron atmosphere and absence of reflection, the emission current is given by—

$$i = \frac{N^3 e^2}{2R} \sqrt{\frac{k}{2m}}\, C^2\, T^2 \exp(\mu/RT),$$

which is of the form of the second Richardson equation (μ is negative at all temperatures). The thermodynamic potential μ is then shown to equal minus the work necessary to extract from the electrode a mol of electrons at constant temperature.

LAUE, *Entropy Constant of Thermionic Electrons* (1.171), working on the analogy between an ideal gas and the electron atmosphere, gives reason for thinking C a universal constant, and in another paper published simultaneously, the entropy constant of electron gas is evaluated, and from this a value of C is derived. The corresponding constant in the formula for the saturation current i becomes

$$A = \frac{2\pi\, e\, m\, k^2}{h^3}$$

so that

$$i = \frac{2\pi\, e\, m\, k^2}{h^3}\, T^2 \exp(\mu/kT).$$

The current i and the charge e are here given in Lorentzian electrostatic units. Reverting to ordinary E.S.U. the outward form of the result remains unchanged, and the numerical value of the constant A is given by—

$$A = \frac{2\pi \times 9{\cdot}02 \times 10^{-28} \times 4{\cdot}774 \times 10^{-10} \times 1{\cdot}372^2 \times 10^{-32}}{6{\cdot}55^3 \times 10^{-81}}$$
$$= 1{\cdot}80 \times 10^{11} \text{ e.s.u. /cm}^2 \text{ deg}^2.$$
$$= 60{\cdot}2 \text{ amps. /cm.}^2 \text{ deg.}^2$$

In the above investigations, LAUE definitely excludes the image force potential from the fundamental equations and concludes therefore that his results apply only to sufficiently high temperatures and, as far as electron distribution is concerned, only beyond a certain small distance from the emitting surfaces. LAUE, *Rôle of the Image Force in the Thermodynamics of Thermionic Electrons* (1.172), in a later paper shows, however, that these restrictions were unnecessary, that the image force plays no special part and that the corresponding potential must be included in the ordinary potential function ϕ. The assumption underlying the whole of the investigation is that the space charges are sufficiently dense to give rise to a potential function which does not vary with time; no further assumption is necessary to include the "image forces."

LAUE, *Conditions for the Validity of the Electron Gas Concept* (1.173), returned to the question of the conditions under which an

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electron atmosphere may be treated as a perfect gas, in a paper of 1919. He shows that for a totally enclosed space, the walls consisting of emitting material and the whole maintained at constant temperature T, the necessary and sufficient condition that the electron density shall not vary throughout the enclosure by more than 1 per cent. is that

$$\frac{2\pi e}{kT}\,\rho\, l^2 < 1,$$

where l is the linear dimension of the enclosure and ρ is the density of the electrons at the walls. For tungsten at a temperature of 2,400° absolute this implies $l \leq 10^{-4}$ cms. For higher temperatures as the formula shows, l must be even smaller (for ρ increases with T much faster than the first power of T). For the numerical case just mentioned with $l = 10^{-4}$ the total number of electrons in the enclosure is only 5. Although LAUE criticises for these reasons the application of perfect gas laws to the electron atmosphere in equilibrium with an emitter, he points out that the final emission formula derived by their aid is the same as that deduced by him using more rigorous methods.

The position taken up by LAUE with respect to the applicability of simple thermodynamic methods to the electron emission process, and with respect to the image force, was challenged by SCHOTTKY, *Electron Vapour Pressure and Clausius-Clapeyron Equation* (1.174). SCHOTTKY claims to show in the first place that Laue's formula

$$\rho_s \propto T^{3/2} \exp(\mu/RT)$$

is identical in substance with the expression deduced by a naive application of the Clausius-Clapeyron equation leading to

$$\rho_s \propto \exp\{w/RT^2\}.$$

A fallacy in Schottky's argument is, however, detected by LAUE, *Applicability of the Clausius-Clapeyron Equation to Thermionic Electrons* (1.175). Schottky's remarks on the image force, *Further Remarks on the Electron Vapour Problem* (1.176), constitute in the main an amplification of his earlier paper (1.130).

The relations amongst thermo-electric, volta, and thermionic effects deduced by RICHARDSON are criticised by BRIDGMAN, *A Critical Discussion of the Volta, Thermo-electric and Thermionic Effects* (p.180), on much the same lines as by SCHOTTKY. BRIDGMAN emphasises the necessity of distinguishing between the heat absorbed by the system due to the evaporation of an electron keeping the surface charge constant, and the heat absorbed when a corresponding positive charge is produced on the emitting surface. The difference of the two latent heats is put equal by BRIDGMAN to a surface heat of charging P_s. The Richardson formula

$$\frac{\partial w}{\partial T} + e\sigma = \frac{k}{\gamma - 1}$$

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is then replaced by

$$\frac{d\eta_\rho}{dT} + e\sigma - \frac{eP_s}{T} = \frac{k\gamma}{\gamma - 1},$$

where η_ρ is the latent heat which is absorbed when evaporation takes place at constant surface charge. (Richardson's w corresponds to the internal latent heat and Bridgman's η_ρ is the total heat, so that $\eta_\rho = w + kT$. The only difference between the formulæ is, therefore, the term involving the surface heat.) BRIDGMAN points out that the existence of a surface heat had already been anticipated by KELVIN and LORENTZ.

The expression obtained by LAUE for the emission current $i = AT^2 \exp(\mu/RT)$, involves the free energy of the electron. SCHOTTKY, *Ion Equilibrium and Contact Potentials* (1.190), makes the suggestion that μ is separable into two parts, one due to the actions of the immediate surroundings of the electron inside the solid, and another proportional to the potential difference between the inside of the metal and the critical plane outside the surface at which emission occurs (see 1.130).

A discussion by TOLMAN, *The Entropy of Electron Gas* (1.200), of the entropy of electron gas, assuming this to be given by the formula

$$S = \frac{5}{2} R \log T - R \log p + (3/2) R \log M + S_1$$

indicates satisfactory agreement between the entropy computed from the emissions of different metals and from their work functions derived by the heat absorption method.

Following on the work of LAUE and TOLMAN, DUSHMAN, *A General Relation for Electron Emission from Metals* (1.201), *Electron Emission from Metals as a Function of Temperature* (1.202), gives a slightly different method of deriving the emission equation with A equal to a universal constant, which brings out clearly the nature of the assumptions involved. Putting for the electron evaporation from a metal, L = heat of vaporisation per gram-molecular weight at the absolute temperature T, R = gas constant per gram-molecular weight = 1·987 cal./deg., p = vapour pressure at temperature T, the second law of thermodynamics gives

$$L = RT^2 \frac{d \log p}{dT}.$$

L can be written

$$L = L_o + \int_0^T C_p dT - \int_0^T c_p dT,$$

where C_p and c_p denote the specific heats at constant pressure of the electrons, as vapour and in the metal, respectively. By integration it follows that

$$\log p = -\frac{L_o}{RT} + \frac{1}{R}\int^T \frac{dT}{T^2}\int_0^T C_p dT - \frac{1}{R}\int^T \frac{dT}{T^2}\int_0^T c_p dT + i.$$

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The value of the integration constant i was shown by SACKUR to be expressible in the case of monatomic gases by $i = i_o + {}^3/_2 \log_e M$, where M is the molecular weight. SACKUR, *Die Anwendung der kinetischen Theorie de Gasl auf chemische Problems* (1.207), and TETRODE, *Die chemische Konstante der Gasl und das elementare Wirkungsquantum* (1.208), give for i_o the expression

$$\log_e \left[\frac{(2\pi)^{3/2} k^{5/2}}{N^{3/2} h^3} \right].$$

The essential point in the argument is now that c_p is negligible for all temperatures under consideration. This means that the electrons within the metal are not present as an ordinary gas and make only an inappreciable contribution to the specific heat of the metal. On these assumptions the thermionic emission formula runs

$$\text{emission current} = \left(\frac{2\pi k^2 me}{h^3} \right) T^2 \exp(-L_o/RT).$$

$$\text{The universal constant } A = \frac{2\pi k^2 me}{h^3} = 60{\cdot}2 \text{ amps./cm.}^2 \text{ deg.}^2$$

$$= 1{\cdot}80 \cdot 10^{11} \text{ e.s.u./cm.}^2 \text{ deg.}^2$$

is the same as that already obtained by LAUE. DUSHMAN points out that by adopting the theory of LEWIS, GIBSON and LATIMER, *A Revision of the Entropies of the Elements* (1·209), another value for i is obtained, leading to

$$A = \frac{2^{5/2} \pi^{9/2} \varepsilon^{-5/2}}{15} \cdot \frac{k^2 me}{h^3} = 1{\cdot}53 \cdot 10^{11} \text{ e.s.u./cm.}^2 \text{ deg.}^2$$

$$= 51{\cdot}2 \text{ amps./cm.}^2 \text{ deg.}^2$$

ε = base of natural logarithms.

RICHARDSON (1.121) had already conjectured from quantum theory considerations that A would be a universal constant and gave the value $A = 1{\cdot}47 \cdot 10^{10}$ e.s.u./cm.^{2} deg.^{2} An experimental differentiation between the Sackur-Tetrode and Lewis-Gibson-Latimer A values is regarded by DUSHMAN as being outside the limits of the observational accuracy obtainable. A discussion of the above considerations in more popular form was given by DUSHMAN, *Theory of Electron Emission* (1.203), in a paper in 1923.

The historical genesis of the emission equation $AT^2 \exp(-b/T)$ is dealt with by RICHARDSON, *Electron Emission from Metals as a Function of Temperature* (1.204), who points out that he gave the result in 1911 and stated in 1915 that A was a universal constant. RICHARDSON also gave the correct value k^2me/h^3 for the dimensional factor entering into the universal constant, in a paper of 1914.

Commenting on the formula $i = AT^2 \exp(-b/T)$, PONTREMOLI, *Thermionic Emission* (1.205), shows that it can be obtained directly with the Sackur-Tetrode value of the constant A, viz., $2\pi mek^2/h^3$,

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by applying to thermionic emission, Saha's equation for the reaction isobar in the case of an ionisation reaction

$$M \rightleftarrows M_1 + el - U,$$

where U is the heat of ionisation of a molecule ($U = eV$ where V is the ionisation potential of the molecule). The final formula is

$$i = \frac{2\pi mek^2}{h^3} T^2 \exp(-eV/kT).$$

The way in which the chemical constant is introduced into the thermodynamics of electron emission and its evaluation, using the quantum theory, are examined by H. A. WILSON, *The Theory of Thermionics* (1.206). The simplifying assumptions which must be made in order to obtain the Laue-Dushman result are carefully indicated.

The complicated case of the simultaneous emission of positive ions, electrons and neutral atoms was treated by LAUE, *Theory of the Positive Ions and Electrons Emitted by Glowing Metals* (1.210), using the methods already employed by him in his earlier work. The paper included a further discussion of the " image force " and the manner in which it has to be allowed for.

In a further paper LAUE and SEN, *Calculation of the Potential Fall in the Ion and Electron Gas Emitted by Glowing Metals* (1.211), work out numerically the particular case in which a plane metal surface at $x = o$ is in equilibrium with the uncharged vapour at $x = \infty$. For this case there is a steady drop (or rise) of potential in passing from the metal to infinity, the amount of which is proportional to the logarithm of γ, the ratio of the equilibrium densities of electrons and positive ions at the surface. The distance x_s from the surface at which a given fraction S of this potential drop (or rise) has already occurred is little affected by the particular value of γ, although depending on the temperature. For $\gamma = 1/10$, S = 99 per cent., the critical distance x_s for potassium at 500° C. is computed to be 80 cm., whilst for the same metal at 350° C. it equals 93 metres.

The applications of thermodynamics to electron atmospheres in equilibrium with matter were reconsidered by RICHARDSON, *Thermodynamics of Electron Emission* (1.220), in 1924, treating the electrons as forming, outside the metal, an ideal gas. A more precise specification of the quantities appearing in the formulæ is given than in previous papers, and the results obtained may with advantage be summarised in some detail.

The circuital system employed is the quasi-equilibrium one already mentioned, in which two solid emitters at different temperatures each in equilibrium with its own electron atmosphere

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are joined by means of an insulated and very fine wire. The symbols have the following meanings :—

p = pressure of electrons in equilibrium with the substance.

V = potential just outside substance.

w = internal latent heat of evaporation of electrons per electron.

γ = ratio of specific heats of electron gas.

σ = reversible heat absorbed when unit quantity of electricity moves against unit rise of temperature in the metal.

The principle used is $\int \frac{dQ}{T} = 0$ for a cycle in which a small charge is passed round the circuit. Applying this cycle to the case in which both emitters are of the same substance, it is shown that for a simple emitter which does not possess transition points in the temperature range considered

$$p = \exp \frac{C_1}{k} \cdot T^{\gamma/\gamma-1} \exp \left\{ -\frac{w}{kT} - \int_0^T \frac{e\sigma}{kT} dT \right\}.$$

The constant C_1 here is considered by RICHARDSON to be a universal constant, on the following grounds. If the thermodynamic argument be applied when the two emitters are of different materials the equation obtained runs

$$w_2/T_2 - w_1/T_1 + k \log p_2/p_1 - \frac{k\gamma}{\gamma - 1} \log T_2/T_1$$

$$+ e \int_{T_0}^{T_2} \frac{\sigma_2}{T} dT + e \int_{T_1}^{T_0} \frac{\sigma_1}{T} dT + \frac{eP_{12}}{T_0} = 0 \quad . \quad . \quad . \quad . \quad . \quad (1)$$

where the emitter (1) is at temperature T_1, the emitter (2) is at temperature T_2, and the temperature of the junction between the two emitters (situated in connecting wire) is T_0 ($T_2 > T_0 > T_1$). P_{12} is the reversible heat absorbed at the junction on passage of unit quantity of electricity; in fact, the Peltier coefficient. A further relation is obtained by passing a small charge round a simple thermo-electric circuit with one junction at T_0 and the other at the absolute zero.

Applying again $\int \frac{dQ}{T} = 0$ it follows that

$$\int_0^{T_0} \frac{\sigma_2}{T} dT - \int_0^{T_0} \frac{\sigma_1}{T} dT - \frac{P_{12}}{T} + \left(\frac{P_{12}}{T}\right)_{T=0} = 0 \quad . \quad . \quad . \quad (2).$$

If $\left(\frac{P_{12}}{T}\right)_{T=0} = 0$ for all pairs of substances, then eliminating $\frac{P_{12}}{T_0}$

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between the equations (1) and (2) the relation obtained is

$$\frac{w_2}{T_2} + k \log p_2 - \frac{k\gamma}{\gamma - 1} \log T_2 + e \int_0^{T_2} \frac{\sigma_2}{T} dT$$
$$= \frac{w_1}{T_1} + k \log p_1 - \frac{k\gamma}{\gamma - 1} \log T_1 + e \int_0^{T_1} \frac{\sigma_1}{T} dT.$$

As each side contains only quantities referring to one substance it follows that

$$\frac{w}{T} + k \log p - \frac{k\gamma}{\gamma - 1} \log T + e \int_0^T \frac{\sigma}{T} dT = C_1$$

is a universal constant. For the validity of the relation $\left(\frac{P_{12}}{T}\right)_{T=0} = 0$ RICHARDSON advances several arguments, among which may be mentioned :—

(1) According to experiments at low temperatures the thermo-electric powers of various pairs of metals fall away rapidly as the absolute zero is approached.

(2) The Nernst Heat Theorem requires the thermo-electric power at very low temperatures to become infinitely small, of the same order as the specific heat.

The argument that C_1 is a universal constant is restricted to metallic substances, although, of course, its invariance with temperature holds good generally.

The saturation current i corresponding to the equilibrium electronic pressure p is obtained simply from

$$i = en \sqrt{\frac{kT}{2\pi m}}$$

$$p = nkT = \exp (C_1/k) \, . \, T^{\gamma/\gamma - 1} \exp \left\{ - \frac{w}{kT} - e \int_0^T \frac{\sigma}{kT} dT \right\}$$

$$\gamma = 5/3.$$

We have, in fact, assuming the absence of reflection of electrons by the surface

$$i = C_2 T^2 \exp \left\{ - \frac{w}{kT} - \int_0^T \frac{e\sigma}{kT} dT \right\}.$$

The integral in the exponential is too small to be of importance with the present accuracy of measurement; w may vary with temperature. w was defined as the internal latent heat of evaporation per electron, i.e., the excess of the average energy of an electron outside over that of an electron inside the emitter, under equilibrium conditions.

Thus $w = \frac{3}{2} kT + \xi - \eta$, where ξ is the excess of the potential energy of an electron outside over that of one inside, and η is the average kinetic energy of an electron inside the emitter. The

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cooling effect per electron, L, due to the emission is, however, given by the average kinetic energy of an electron leaving the surface, $2k\mathrm{T}$, added to ξ, less $f(\eta)$, the average kinetic energy of electrons which carry the emission current, inside the emitter.

Thus $\mathrm{L} = 2k\mathrm{T} + \xi - f(\eta)$. The values of η and $f(\eta)$ will depend on the view held as to the condition of electrons in conductors. For the classical theory, assuming the electrons inside to approximate to a perfect gas

$$\eta = 3/2\ k\mathrm{T}$$
$$f(\eta) = 2k\mathrm{T}$$

so that $\mathrm{L} = w$.

On the non-classical type of theory η and $f(\eta)$ are either exactly or approximately independent of temperatures and

$$\mathrm{L} = w + \tfrac{1}{2}\,k\mathrm{T} + \eta - f(\eta).$$

It is reasonable to expect also $\eta = f(\eta)$.

Thus for non-classical theories, which are the only ones likely to be valid, the cooling effect should exceed the internal latent heat of evaporation. As the experimental data determined by DAVISSON and GERMER show, for tungsten the measured L exceeds the measured w by the equivalent of 0·132 volts, which is equal, within the limits of experimental accuracy, to $\frac{1}{2}\,k\mathrm{T}$ (expressed in volts). Thus the evidence is in favour of the non-classical theories which make the kinetic energy of the conducting electrons in metals independent of temperature.

RICHARDSON also considers the criticisms by SCHOTTKY and BRIDGMAN of his equation

$$\frac{\partial w}{\partial \mathrm{T}} = \frac{k}{\gamma - 1} - e\sigma.$$

RICHARDSON agrees that strictly the equation should run

$$\phi = w - \mathrm{T}\frac{\partial w}{\partial \mathrm{T}} + \frac{k}{\gamma - 1}\,\mathrm{T} - e\sigma\mathrm{T},$$

where $\dfrac{w - \phi}{e}$ equals the surface heat of charging P_s.

Reasons are given, however, for thinking the surface heat negligible to the degree of experimental accuracy with which w is determinate. An approximate estimate indicates that the upper limit to the value of P_s for tungsten at 2,300° K. is 5 per cent. of w.

The mechanism of thermionic emission is discussed in detail by RICHARDSON, *Thermionic Emission from Systems with Multiple Thresholds* (1.221), for the case when electrons may originate from several thresholds (i.e. the work done by an electron in leaving the metal, having one of a number of distinct values). The problem is approached from three different standpoints: (*a*) classical statistical mechanics, (*b*) chemical dynamics, (*c*) the laws of photo-electric action. All three methods lead to difficulties, but RICHARDSON

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lays down a set of laws governing the emission, which are at any rate not in disagreement with known facts or well-accepted principles.

DAVISSON, *A Note on the Thermo-dynamics of Thermionic Emission* (1.229), deduces the equation

$$n = \mathrm{A} \exp \left\{ \int_0^{\mathrm{T}} \frac{w d\mathrm{T}}{k\mathrm{T}^2} \right\}$$

for the concentration of electrons just outside an emitter in the equilibrium state, by applying thermodynamics to a system consisting of a parallel plate condenser, one plate of which is the emitter and the other is a perfect reflector not emitting electrons. Allowance is made in the same way as in Laue's work, for the variation of electron concentration due to the field between the plates. The present proof of the formula is free from the conditions underlying the earlier derivations (except Laue's), which required that the temperature should be sufficiently low for variation of electron concentration due to space charge effect to be negligible.

A theory of metallic conduction was developed by WATERMAN, *The Variation of Thermionic Emission with Temperature and the Concentration of Free Electrons within Conductors* (1.230), in a paper of 1924, in which bound and free electrons in the metal are contemplated, the energies required at the absolute zero to separate a bound and a free electron from the metal being respectively ϕ_0 and ψ_0. The atoms of the metal are assumed to liberate electrons in accordance with the reaction equation, atom $\rightleftarrows$ positive ion + ν electrons. It is shown that the number of electrons per unit volume in equilibrium with the hot metal at temperature T is given by

$$n = \exp\left(\frac{\nu j}{\nu + 1}\right) \cdot (\nu \mathrm{N})^{1/\nu + 1} \cdot \mathrm{T}^{3\nu/2(\nu+1)} \cdot \exp\left\{ - \frac{\nu (\phi_0 + \psi_0)}{(\nu + 1) k\mathrm{T}} \right\},$$

where N is the number of atoms per unit volume in the conductor and j is the chemical constant of the electrons. Writing the above expression in the form $n = \mathrm{CT}\alpha \exp(-\beta/\mathrm{T})$ it is seen that C and α are dependent on the number of electrons ν involved in the ionisation of the metal atoms.

ROY, *On the Law and Mechanism of the Emission of Electrons from Hot Bodies* (1.234), discussing Dushman's derivation of the emission equation, gives an explanation of the negligible contribution to the specific heat of the electrons within the metal, assuming them to be arranged in space lattices just as atoms in a crystal, and further assuming that the electrons emitted under thermal impulse are identical with the photo-electrons.

Roy's final formula is $i = \frac{2\pi mek^2}{h^3} \mathrm{T}^2 \exp\left(-\frac{h\nu_0}{k\mathrm{T}}\right)$, where ν_0 is the threshold frequency of the photo-electric effect at the absolute zero.

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A constructive theory of thermionic emission is outlined by KINGDON, *A Mechanism for Thermal Electron Emission* (1.238). It is assumed that the atoms in the emitter surface oscillate with amplitudes distributed in accordance with Maxwell's distribution law. A collision is said to occur when the amplitude equals the distance between adjacent atoms, each collision resulting in the ejection of an electron. This theory seems to lead in the case of tungsten to an expression for the emission current which agrees with experiment to within 10 per cent. over the entire temperature range available and which contains no arbitrarily assignable constant.

A general investigation of the emission equation was carried out by RASCHEVSKY, *Evaporation of Electrons* (1.240), in 1925, in which Fowler's new statistical methods were employed. RASCHEVSKY assumes (1) that the emission of the electrons can be regarded as an evaporation process, (2) that the evaporated electrons behave as an ideal gas, (3) that energy interchange between electrons and atoms in the metal is negligible. It is shown (1) that on statistical and thermodynamic grounds the emission formula has the general form $i = AT^a \exp(-b/T)$ and that a equals 2 if, and only if, the energy of the interior electrons is regarded as independent of temperature; this holds good in both classical and quantum theory treatments; (2) that the assumption of free electrons inside the body leads necessarily to $a = \frac{1}{2}$, again for both classical and quantum methods; (3) that if the electrons are assumed to be arranged in a space lattice, then on the classical theory $a = -1$ and

$$i = \frac{2\pi m \nu^3}{k} e\, T^{-1} \exp(-w/kT),$$

where ν is the vibration frequency of the electron lattice and w is the work required to transport an electron from within the body to the surrounding space; (4) that on the quantum theory (assuming quantised motion, all electron orbits in the metal having the same quantum number) $a = 2$ and

$$i = 2n^1 \pi m h^2 T^2 \exp\left\{-\frac{w - E}{kT}\right\},$$

where n^1 is the number of electrons per unit volume in the metal and E is the energy in each electron orbit.

Similar results are deduced by RASCHEVSKY, *On the Kinetic Theory of the Thermionic Effect* (1.241), using slightly different methods, in an American publication.

RASCHEVSKY regarded his third assumption (above) as unsatisfactory and only capable of giving approximations. For the statistical handling of the problem, the evaporation of electrons and of atoms cannot be dealt with separately. To take into account the connexion between the two phenomena, RASCHEVSKY, *Thermionic Effect from the Standpoint of the Phase Rule* (1.242) considers the system consisting of two components, electrons and positive ions in two phases, solid and gaseous, when equilibrium is

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attained. An immediate inference is that the thermionic emission depends not only on the temperature, but also on the external pressure. In order to carry out the analysis some assumption must be made as to the thermodynamic potential of the emitter, and RASCHEVSKY derives this on the plausible hypothesis that the emitter may be treated as a solid solution of electrons in the positive ion lattice. A general formula for the emission current is obtained of the form $ABT^2 \exp(-b/kT)$, where A is a universal constant, but B is a constant depending on the vapour pressure law for the neutral atoms and positive ions. RASCHEVSKY also deduces a relation between the electron emission, vapour pressure and specific heat of a body.

In SCHOTTKY's *Evaporation of Electrons* (1.243), investigation of the emission formula is based on the expression

$$\alpha T^2 \exp\{-[\mu]/kT\}$$

$$\alpha = \frac{2\pi k^2 me}{h^3}, \quad [\mu] = \mu + F(\phi_g - \Phi),$$

where μ = the thermo-dynamic potential of a mol of electrons inside the metal, Φ = average electro-static potential inside the metal, ϕ_g = the electro-static potential just outside the metal at the position of the potential minimum as determined by image force, space charge, and external field. F is a numerical constant. All the features of the emission characteristic are included in this result and the problem is to determine μ and $(\phi_g - \Phi)$. The evaluation of the former quantity is a statistical problem concerning conditions in the body of the metal. That of the latter depends, on the other hand, on the arrangement of the charges in the surface. SCHOTTKY concludes that the controlling factor in determining μ is "the change of specific heat of the metal with change in the number of electrons per unit volume contained in it" rather than "the specific heat of the electrons within the metal," the two quantities becoming identical only if the kinetic and potential energies of the electrons are to be computed without reference to a possible interaction with the other constituents of the metal. With respect to $(\phi_g - \Phi)$ it appears still uncertain whether there exists in the surface an electric double layer opposing the emission of electrons.

The new statistical method of FOWLER and DARWIN was applied by ROY, *On the Statistical Theory of Emission of Electrons from Hot Bodies* (1.245), to a system consisting of atom and electron lattices to derive the emission formula

$$i = \frac{2\pi mek^2}{h^3 K_2(T)} T^2 \exp(-w/kT)$$

where $K_2(T)$ is determined by the energy content of the electron lattice. The excess, at high temperatures, of specific heats of metals over the Dulong and Petit value 3R is attributed by ROY to the contribution of the electrons, and by an approximate calculation

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it is shown that $K_2(T)$ cannot differ much from unity even up to the melting point of the metal. ROY also derives the law

$$i = \frac{2\pi.mek^2}{h^3} T^2 \exp(-w/kT)$$

by treating thermionic emission as a form of total photo-electric emission throughout the metal, and using the theory of unit mechanism developed by KRAMERS and MILNE.

The notion that electrons are distributed in the crystalline metal in a space lattice arrangement was employed by WEIGLE, *The Heat of Evaporation of Electrons* (1.250), *Heat of Evaporation of Electrons of Calcium* (1.251), *Lattice Energy and the Heat of Evaporation of Electrons in Calcium* (1.252), to evaluate their heat of evaporation by the thermo-chemical methods used by BORN. The necessary data (heat of evaporation of the neutral atoms, heat of ionisation of the gaseous atoms, the lattice energy, etc.) are known for the alkali metals Na, K, Rb, and Cs and WEIGLE shows that the heat of evaporation of the electrons computed in this way agrees well with the corresponding constant in the thermionic emission equations for the respective metals. Further calculations for Ca, Sr, Ba, Ag, and Cu, are given in later communications.

REICHENSTEIN, *Phenomenology of the Richardson Effect* (1.260), after a general discussion of thermionic emission concludes that the assumption underlying all derivations of Richardson's equation, namely, the very rapid setting up of the equilibrium between the metal and the electron vapour, is not in accord with experimental facts, and that the inertial effects involved call for a " phenomenological " description of electron emission. As the result of an investigation on these lines, REICHENSTEIN anticipates a number of effects, among which may be mentioned that the saturation current at low temperatures, say 100° C., will be much smaller than that computed by extrapolation of the emission equation established at temperatures between 2,000° C. and 3,000° C.

RASCHEVSKY, *Principles of the Thermionic Effect* (1.269) compares his earlier work, which was subjected to criticism by SCHOTTKY, with Schottky's own investigation. The effect of pressure on thermionic emission is worked out more generally.

A thermodynamic cycle is used by BRIDGMAN, *The Universal Constant of Thermionic Emission* (1.270), in investigating the emission equation, which assumes the legitimacy of treating the positive charge left on the surface when electrons are evaporated, as a substance possessing a specific heat. If the surface charge is assumed to behave as a condensed system, so that its energy vanishes at the absolute zero, the condition that the emission equation shall have the form

$$i = AT^2 e^{-b/T} \text{ where } A = \frac{2\pi mek^2}{h^3}$$

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is a universal constant, becomes, specific heat of neutral metal = specific heat of surface charge. In terms of the surface heat P_s this condition takes the form

$$\sigma + \frac{\partial P_s}{\partial T} - \frac{P_s}{T} = o.$$

The fact that emitters with oxide-coated surfaces do not lead to the universal value of the constant A is then attributed by BRIDGMAN to differences between the specific heats of the neutral substance and the surface charge left by electron emission.

RASCHEVSKY, *Theory of the Thermionic Effect* (1.271), arrives at a similar result to Bridgman's for the condition of universality of the constant A, using a different method. RASCHEVSKY argues that for pure metals in which the electrons are arranged in a space lattice or, according to a new conception of FRENKEL, are " bound " not to individual atoms but to the crystal as a whole, the removal of an electron will have comparatively little effect on the heat motion of the neighbouring atom ions. The specific heat of neutral metal and positive charge will be practically the same and A will be a universal constant. In the case of a heteropolar compound X + Y, however, in which the X atom has lost an electron taken over by the Y atom, it is probably just this valency electron which is emitted thermionically. The atom Y is thereby changed from a negative ion to a neutral atom, a transformation which is likely to have a considerable effect on its motion. In this case the condition for universality of A is not obeyed. Oxides provide examples of such heteropolar compounds and experiment confirms that the corresponding values of the constant A vary over a wide range. An explanation of the observed deviation from the universal A value for emitters consisting of a very thin (monatomic) layer of one substance adsorbed on the surface of another, is developed on the same lines. In the same paper, a revised treatment is given of the variation of thermionic emission with pressure, in which it is shown that the effect is vanishingly small.

The question as to whether in the thermodynamics of the thermionic effect the thermodynamic functions should be taken for the interior of the emitter or for its surface, is discussed by RASCHEVSKY, *Theory of the Thermionic Effect II* (1.272). In the same place, a further investigation is made of the universality condition for A.

Commenting on Michel and Spanner's experimental results SCHOTTKY, *Cooling Effect for Oxide Cathodes* (1.279), showed that the cooling effect for a pure substance when covered with a thin layer of another substance will vary to a first approximation by the same amount as the electron exit work occurring in the emission formula.

HALL, *Photo-electric Emission, Thermionic Effect and Peltier Effect* (1.280), discusses the theoretical interpretation of Davisson and Germer's measurements of the cooling effect from the point of view of the dual theory of conduction. In a later paper, *Thermionic*

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Emission and the "Universal Constant A" (1.281), he applies this theory in order to ascertain if the constant A in the emission formula, should be a universal constant. He concludes that it will *not* be the same for all metals.

The question of the surface heat of charging was further examined by TONKS and LANGMUIR, *On the Surface Heat of Charging* (1.290), who attempt to estimate its value (*a*) for pure metal surfaces; (*b*) for other surfaces. For two different surfaces the difference of the surface heats is shown to be given by $kT \log (A_1/A_2) + eP_{12}$ where A_1 and A_2 are derived from the emission equations

$$\{i = AT^2 \exp(-b/T)\}$$

of the respective surfaces and P_{12} is the Peltier heat when the surfaces are in contact. For pure metals P_{12} is small and A_1 and A_2 are nearly equal to 60.2. It follows that the surface heats of charging are the same for pure metals (to within a few per cent. of the work function). By a comparison of the work functions for tungsten determined by DAVISSON and GERMER, using (*a*) the cooling effect; (*b*) the temperature variation method, it is further shown that the absolute magnitude of the surface heat for all pure metals must amount to only a small fraction of the work function. Similar methods, applied to thoriated and coated filaments, lead to difficulties which are perhaps to be met by treating the emitting surface as "patchy."

DAVISSON and GERMER, *Note on the Thermionic Work Function of Tungsten* (1.291), discussing the relationship between the work function determined by the two methods, show that, contrary to Tonks and Langmuir's conclusion, the difference between the two quantities does not involve the surface heat of charging, but only the average kinetic energy of the conduction electrons within the metal and the average kinetic energy transported by these electrons. The small difference between the observed values of the two work functions is interpreted as evidence that the conduction electrons within the filament do not possess normal thermal energies.

A partial refutation of Davisson and Germer's criticism is given by TONKS, *Note on Thermionic Emission* (1.292).

A number of results bearing on the theory of thermionic emission were derived by BRIDGMAN, *General Considerations on the Photo-electric Effect* (1.300), in a discussion of the photo-electric effect. BRIDGMAN concludes that the following three quantities must be universal constants: (1) the difference between the photo-electric and thermionic work functions, (2) the difference between the specific heat of the metal and of a surface charge, (3) the difference between the entropy of the metal and of a surface charge, at 0° absolute. Experimental evidence and dimensional considerations lead BRIDGMAN to the result that all these universal constants have the value zero. The explanation of the observed deviation of A from the universal value, in the case of coated emitters, which was put

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forward in an earlier paper, depends on assuming the second of the differences above to be different from zero for coated substances. The present argument disposes of this explanation, and the view is now advanced that the abnormal values indicate that in such cases the system cannot be in equilibrium, and thermodynamic arguments are inapplicable.

BRIDGMAN, *The Photo-electric and Thermionic Emission, A Correction and an Extension* (1.301), in a later paper corrects an error in his previous work, and now finds, not the difference of entropy between the metal and a surface charge, at 0° absolute, but this difference plus h times the temperature coefficient of the photo-electric threshold frequency at 0° absolute, to be equal to a universal constant. It is pointed out then, that in explaining the large departures of A from the universal value, recourse need not be had to the hypothesis that the metal has not reached an equilibrium condition, but rather that the variation of A is to be connected with a temperature coefficient of the photo-electric frequency and a non-vanishing difference between the entropies of the charged and uncharged metals at the absolute zero. The theory is applied in discussing recent measurements by DUBRIDGE of the photo-electric and thermionic work functions.

The conception of evaporation of electrons from a metal used in explaining the thermionic emission from metals is criticised by V. HIPPEL, *Physical Interpretation of Thermal Electron Emission* (1.309), who seeks to replace it by a theory of emission due to temperature ionisation by atom collisions. V. HIPPEL points out certain advantages in developing the emission equation from the latter standpoint.

The difficulties met with in the electron theory of conduction in metals are practically removed by the application by SOMMERFELD, *Electron Theory of Metals based on the Fermi Statistics* (1.400), of the Fermi-Dirac statistics to the electron gas inside the metal. The fundamental point in the theory is that the Pauli principle, that no two electrons are describing orbits specified by the same set of quantum numbers, holds for the electron gas inside a metal. It follows from this assumption that the electron gas within the metal will be "degenerate" (i.e. $pV \neq RT$) and the distribution of velocities of the electrons will be quite different from the Maxwellian distribution. In fact, if dN is the number of electrons per unit volume with velocity components between $u + du$ and u, $v + dv$ and v, $w + dw$ and w, then the new theory shows that

$$dN = g\left(\frac{m}{h}\right)^3 \frac{du\,dv\,dw}{\frac{1}{S}\exp(E/kT) + 1}$$

where $E = \frac{1}{2}m(u^2 + v^2 + w^2)$ is the energy of the electron and $S = \exp(\mu/kT)$, where μ is the thermodynamic potential of

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an electron. If S is much smaller than unity the distribution reduces to

$$dN = g\left(\frac{m}{h}\right)^3 S \exp(-E/kT)\, du\, dv\, dw,$$

and SOMMERFELD shows that for this case $S = \frac{Nh^3}{g}(2\pi mkT)^{-3/2}$

Hence

$$dN = N\left(\frac{m}{2\pi kT}\right)^{3/2} \exp(-E/kT),$$

which is precisely the Maxwellian distribution. For the electron gas in a metal, however, SOMMERFELD shows that degeneration occurs and S actually is much greater than unity, even when the metal is at the high temperatures required for thermionic emission. The more general expression for the distribution of velocities in the metal must then be taken. The characteristic feature of the new distribution is that the average velocity is practically independent of temperature and is determined principally by the number of electrons per unit volume.

Applying the theory to the thermal emission of electrons, SOMMERFELD evaluates the total number of electrons impinging per unit time on unit area of the metal surface (which is assumed situated in the plane perpendicular to the x axis), with sufficient velocity normal to the surface to overcome a potential energy discontinuity U. Multiplying this number by the electronic charge e, the thermionic emission current is given by

$$i = eg\left(\frac{m}{h}\right)^3 \int \frac{u}{\frac{1}{S}\exp(E/kT) + 1}\, du\, dv\, dw,$$

where the integral is extended from $u = \sqrt{\frac{2U}{m}}$ to $u = \infty$, and from $v = o$, $w = o$ to $v = \infty$, $w = \infty$. Working out the integral, SOMMERFELD finds

$$i = \frac{2\pi meg}{h^3}(kT)^2 S \exp(-U/kT)$$

If the incorrect value $\frac{Nh^3}{g}(2\pi mkT)^{-3/2}$ is assumed for S, i reduces to

$$i = eN\sqrt{\frac{kT}{2\pi m}} \exp(-U/kT),$$

which is Richardson's original result. Applying the new statistics however, S is no longer given by $\frac{Nh^3}{g}(2\pi mkT)^{-3/2}$. By the original

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definition of S, $S = \exp(\mu/kT)$, hence the thermionic emission formula takes the form

$$i = \frac{2\pi mek^2}{h^3} g\, T^2 \exp\left(\frac{\mu - U}{kT}\right)$$

This result agrees in form with Schottky's modification of the original Richardson formula, except that now the factor g enters. g is defined as the quantum weight of a single electron path in the metal. Because of the electron spin, there will be two electrons in each orbit, i.e. $g = 2$. Sommerfeld's theory leads to a value for μ, the thermodynamic potential of the electron, equal to

$$\frac{h^2}{2m}\left(\frac{3N}{4\pi g}\right)^{2/3}.$$

The effective work function $(U - \mu)$ is seen to be the difference between the " external " work done by the electron in overcoming the potential discontinuity at the surface and the " internal " work μ, which may be regarded as the work done *on* the electron by the electron pressure in the metal.

The new electron theory of metals was also employed by FOWLER, *The Restored Electron Theory of Metals and Thermionic Formulæ* (1.410), to obtain the thermionic emission equation. His result is as follows :—

$$i = \frac{2\pi mek^2}{h^3} g\,(1 - r)\, T^2 \exp(-\chi/kT)$$

and differs from Sommerfeld's in the inclusion of the factor $(1 - r)$, where r represents the reflection factor for electrons impinging on the surface. A fuller treatment of the problem, taking into account this reflection factor and computing its value by wave mechanics, was carried out by NORDHEIM, *Theory of the Thermal Emission and Reflection of Electrons by Metals* (1.500).

Taking Sommerfeld's expression for dN, the number of electrons with velocities in the velocity range $du\ dv\ dw$, NORDHEIM derives the number of electrons with kinetic energy normal to the surface lying between $P = \frac{1}{2}mu^2$ and $P + dP = \frac{1}{2}mu^2 + mu\ du$, impinging on unit area in unit time

$$N(P)\, dP = g\left(\frac{m}{h}\right)^3 udu \int_{v=o}^{v=\infty} \int_{w=o}^{w=\infty} \frac{dv\ dw}{\exp\left(\frac{E - \mu}{kT}\right) + 1}$$

Re-arranging the integration over the (v, w) plane by transforming to polar co-ordinates $\rho = \sqrt{v^2 + w^2}$ and we have

$$N(\rho)\, dP = g\left(\frac{m}{h}\right)^3 udu \int_{\rho=o}^{\rho=\infty} \int_{\phi=o}^{\phi=2\pi} \frac{\rho d\rho d\phi}{\exp\left(\frac{P - \mu + \frac{1}{2}m\rho^2}{kT}\right) + 1}$$

$$= \frac{2\pi mgkT}{h^3} \cdot L\left(\frac{P - \mu}{kT}\right) \cdot dP,$$

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where

$$L(a) = \int_{b=o}^{\infty} \frac{db}{\exp(a+b)+1.}$$

The electrons are constrained within the metal by a potential energy discontinuity at the surface which for the case of a clean metal surface may be taken to have the form shown in the Figure

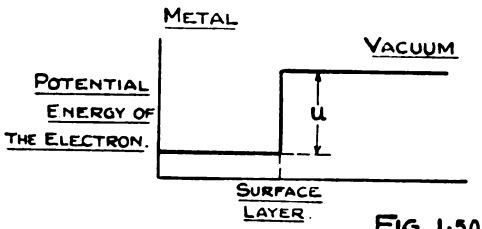

FIG. 1·500 a

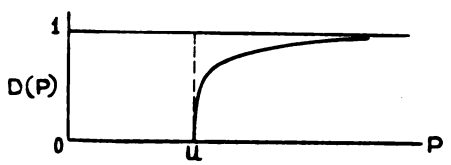

FIG. 1·500 b

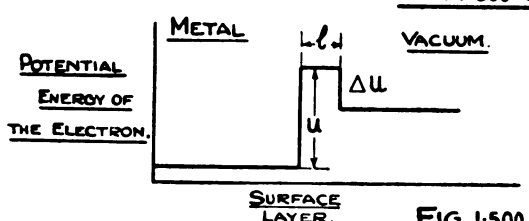

FIG. 1·500 c.

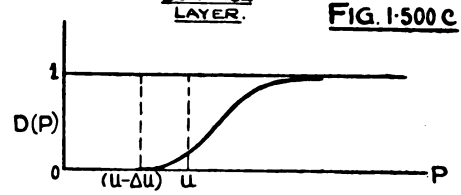

FIG. 1·500 d.

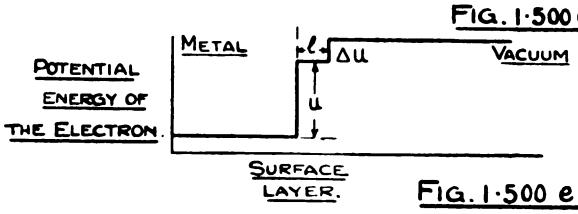

FIG. 1·500 e.

1.500*a*). According to classical ideas every electron with energy P greater than U would escape from the surface. The application of wave mechanics shows, however, that for an electron with energy P greater than U there is only a certain probability, D (P), that it will escape, i.e. some electrons will be reflected.

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Thus the total emission current i will be given by

$$i = e \int_{\rho=u}^{\infty} \mathrm{N}(\mathrm{P})\,\mathrm{D}(\mathrm{P})\,d\mathrm{P}.$$

Inserting the value of N (P) and making suitable approximations, NORDHEIM obtains

$$i = \frac{2\pi mek^2}{h^3} g\,\bar{\mathrm{D}}\,\mathrm{T}^2 \exp\left(-\frac{\mathrm{U}-\mu}{k\mathrm{T}}\right)$$

where

$$\bar{\mathrm{D}} = \int_{P=U}^{\infty} \mathrm{D}(\mathrm{P}) \exp\left(-\frac{\mathrm{P}-\mathrm{U}}{k\mathrm{T}}\right) . d\left(\frac{\mathrm{P}}{k\mathrm{T}}\right)$$

or putting $t = \dfrac{\mathrm{P}-u}{k\mathrm{T}}$

$$\bar{\mathrm{D}} = \int_{t=o}^{\infty} \mathrm{D}(\mathrm{U} + k\mathrm{T}t) \exp(-t)\,dt.$$

Thus $\bar{\mathrm{D}}$ is the average transmission coefficient of all those electrons impinging on the surface, which have sufficient energy P to escape.

NORDHEIM then considers the electron reflection problem by applying Schrödinger's wave equation to the case of electrons impacting on a potential energy discontinuity. For a discontinuity of the form shown in Fig. 1.500*a*, which corresponds to a clean metal surface, D (P) is found in this way to equal

$$4\left(\frac{\mathrm{P}-\mathrm{U}}{\mathrm{P}}\right)^{\frac{1}{2}} \left\{1 + \left(\frac{\mathrm{P}-\mathrm{U}}{\mathrm{P}}\right)^{\frac{1}{2}}\right\}^{-2}$$

(Fig. 1.500*b*). Assuming values of T and U of the order of magnitude met with in thermionic emission, NORDHEIM finds

$$\text{for } \frac{k\mathrm{T}}{\mathrm{U}} = \frac{1}{50},\ \bar{\mathrm{D}} = 0{\cdot}24\,;\quad \text{for } \frac{k\mathrm{T}}{\mathrm{U}} = \frac{1}{100},\ \mathrm{D} = 0{\cdot}18.$$

Thus $\bar{\mathrm{D}}$ is of the order 0·2, and the thermionic emission equation runs, putting for g its value 2,

$$i = 2 \times 0{\cdot}2 \left(\frac{2\pi mek^2}{h^3}\right) \mathrm{T}^2 \exp\left\{-\frac{\mathrm{U}-\mu}{k\mathrm{T}}\right\}$$

DUSHMAN and his collaborators find experimentally for pure tungsten and tantalum the numerical constant in the emission equation to be given approximately by $\left(\dfrac{2\pi mek^2}{h^3}\right)$, which is seen to be about double the value predicted by Nordheim's theory.

Extending the theory to the emission from composite surfaces, NORDHEIM attributes the increased emission of a thoriated tungsten filament to a modification of the potential energy discontinuity

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in some such way as shown in Fig. 1.500*c*. According to classical theory, no electron would escape through the modified surface unless its energy P exceeded U. The wave mechanics treatment shows, however, that electrons with energy P between $U - \Delta U$ and U will have a certain non-zero probability of "percolating" through the hump in the potential energy curve. The mode of variation of D (P) with P is shown in the sketch of Fig. 1.500*d*. The emission equation for the thoriated tungsten surface becomes

$$i = \frac{2\pi mek^2g}{h^3} \bar{D}\, T^2 \exp\left\{-\frac{U - \Delta U - \mu}{kT}\right\}$$

where $$\bar{D} = \int_0^\infty D\,(U - \Delta U + kTt) \exp(-t)\, dt.$$

The value of $\bar{D}$ depends upon ΔU and upon the thickness of the surface layer *l*. It is always less than $\bar{D}$ computed for the clean surface with surface discontinuity U. The observed effect of the thorium layer on the emission is a reduction of the two Richardson constants ϕ and A ($i = AT^2 \exp(-\phi e/kT)$). Nordheim's theory reproduces these changes. Taking Dushman and Ewald's values for fully thoriated tungsten $A = 3$, $\phi = 2{\cdot}63$ volts, and taking ϕ for tungsten as 4·54 volts, it follows that $\Delta U = e\,(4{\cdot}54 - 2{\cdot}63)$, $\bar{D} = \frac{3}{120}$, and from these relations an estimate is made of the thickness of the surface layer *l*. NORDHEIM obtains $l \simeq 3 \times 10^{-8}$, which is of the right order for a mono-molecular layer. For a layer of a more electro-negative material such as oxygen, the variation of potential energy in the neighbourhood of the surface is represented ideally as shown in Fig. 1.500*e* and is discussed on the same lines.

In a further paper, NORDHEIM, *The Effect of the Image Force on the Emission and Reflection of Electrons by Metals* (1.501), modified the potential energy discontinuity at the surface, to take account of the image force. For the modified form (Fig. 1.501*a*), $\bar{D}$ for a clean surface is found to be very nearly unity, only a few per cent. of the electrons eligible for emission being reflected. The thermionic emission equation for pure metals with clean surfaces becomes, putting $\bar{D} = 1$,

$$i = \frac{2\pi mek^2g}{h^3} T^2 \exp\left(-\frac{U - \mu}{kT}\right),$$

or

$$i = 120\, T^2 \exp(-b/T) \text{ amps./cm}^2.$$

Another calculation of the reflection of electrons from composite surfaces on the same lines as Nordheim's was carried out by GEORGESON, *Thermionic Emission through Double Layers* (1.502), who, in order to approximate more closely to actual conditions,

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employed a potential energy distribution as shown in Fig. 1.502*a*. GEORGESON confirmed for this case Nordheim's general conclusions.

A further application of the Sommerfeld theory of electrons in metals was made by FOWLER, *The Photo-electric Threshold Frequency and the Thermionic Work Function* (1.510), who considered the problem of the photo-electric threshold frequency and succeeded in explaining a previously enigmatic result, the sharpness of the cut-off in photo-electric current at the threshold. He showed further that the energy $h\nu_0$ corresponding to this frequency was equal to the thermionic work function, a result which had been long surmised, but for which a satisfactory proof had up to then been lacking.

FOWLER, *The Thermionic Emission Constant A* (1.520), also attacked the problem of the remarkable logarithmic relation between the emission constants A and *b* first recorded by O. W.

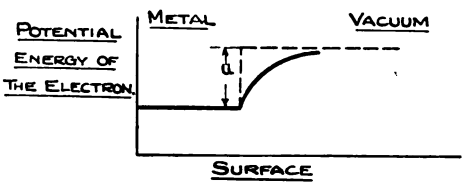

FIG. 1.501*a*

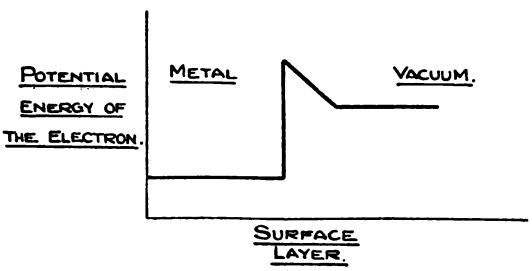

FIG. 1.502*a*

RICHARDSON (2.125) and later drawn attention to by DU BRIDGE (2.315). The relation in question states that changes in the experimental values of the constants A and $b\left(i = AT^2 \exp\left(-\frac{b}{T}\right)\right)$ always go together in such a way that $\log A = \alpha b + \beta$, where α and β are constants ($\alpha > o$), α in particular depending on the type of surface dealt with. FOWLER adopts Nordheim's theory of the thermionic emission equation leading to the result $i = 120\,\overline{D}\,T^2 \exp\left\{-\frac{U-\mu}{kT}\right\}$, where necessarily $\overline{D} \leqslant 1$, and hence $120\,\overline{D} \leqslant 120$. Actually values of A determined experimentally in certain cases far exceed 120 (e.g. DU BRIDGE finds for platinum that A is of the order 10,000). Three possible explanations of the apparent discrepancy might be advanced.

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(1) $\bar{D}$ varies rapidly with temperature so that part of it is included by the experimentalist in the exponential term. The experimental A and b are not then comparable with 120 $\bar{D}$ and $(U - \mu)$.

(2) $(U - \mu)$ varies with temperature as suggested by SCHOTTKY, so that $(U - \mu) = (U - \mu)_0 - akT$, and an extra term exp. a will be included in A determined by the usual experimental analysis.

(3) The areas of emitting surface are wrongly estimated.

Passing over the difficulty in explaining the high absolute values observed for A, FOWLER extends and corrects Nordheim's analysis leading to the value of D for a potential energy discontinuity of the form shown in Fig. 1.500c, and obtains as an approximate formula

$$\bar{D} = 8\pi k \frac{\{kT\,(U - \Delta U)\}^{\frac{1}{2}}}{U} \exp\{-2kl\,(\Delta U)^{\frac{1}{2}}\}$$

where $k^2 = \frac{8\pi^2 m}{h^2}$, provided exp $(-\Delta U/kT)$ is small compared with exp $\{-2\pi\, l\, (\Delta U)^{\frac{1}{2}}\}$ Putting A equal to $120\bar{D}$ and taking logarithms, the equation, obtained runs,

$$\log A = \beta^1 - 2k\,(\Delta U)^{\frac{1}{2}}\, l \text{ where } \beta^1 \text{ is constant.}$$

Inserting in the experimental relation $\log A = \alpha b + \beta$, the value of $b = \frac{1}{k}(U - \Delta U - \mu)$, we obtain

$$\log A = \frac{\alpha}{k}(U - \Delta U - \mu) + \beta$$

or

$$\log A = \beta^1 - \frac{\alpha}{k}\Delta U$$

Changes in ΔU and l are thus seen to produce theoretically changes of A of approximately the type obtained experimentally. Applying the theoretical formula to Du Bridge's results for platinum, FOWLER finds agreement if l is taken to be 10^{-7} cm., a rather thick surface layer, but still a reasonable value.

If exp $\left(-\frac{\Delta U}{kT}\right)$ is not small compared with exp $\{-2kl\,(\Delta U)^{\frac{1}{2}}\}$ $\bar{D}$ is no longer approximately temperature independent and the approximate formula for $\bar{D}$ breaks down. FOWLER proceeds to discuss the possibilities of explaining the large absolute values of A, in terms of the theory, and makes several suggestions.

A connected account of the researches of NORDHEIM and FOWLER on the application of the Fermi-Diræ statistics and wave mechanics theory of reflection at potential barriers to the problems of thermionic emission from clean and composite surfaces, " cold " electron emission and the photo-electric effect, is contained in a comprehensive paper by NORDHEIM, *Theory of the Electron Emission of Metals* (1.525).

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The linear relation between log A and b was discussed by ZWIKKER, *Effect of Surface Layers on the Thermionic Emission of Metals* (1.530), working on different lines from those followed by FOWLER. ZWIKKER writes the relation in the form $b = c \log_e A + d$ and notes that it holds for surface layers of various elements on a given base metal (tungsten, platinum, zirconium and hafnium). For these cases c turns out to have values equal to about 3,000° K, which is of the same order as the absolute temperature of the emitting substances. If T_0 represents the average temperature at which measurements were taken, it is a tolerable approximation to write

$$b = 2T_0 \log_e A + d$$

Denoting by $A_0 \left(= \frac{2\pi mek^2}{h^3}\right)$ and b_0 the Richardson constants for the clean metal surface $b = b_0 + 2\ T_0 \log_e \frac{A}{A_0}$. ZWIKKER assumes with SCHOTTKY that the changes in the experimentally determined A correspond to a temperature dependence of the exit work ϕ^x, as shown by the following equations

$$\frac{e}{k}\phi^x = b^x = b + \beta T$$

$$i = A_0 T^2 \exp\left(-\frac{b + \beta T}{T}\right) = A_0 \exp(-\beta).\ T^2 \exp\left(-\frac{b}{T}\right).$$

$$A = A_0 \exp(-\beta).$$

ZWIKKER regards the exit work ϕ^x of the contaminated surface as the sum of the exit work ϕ_0 of the clean surface and a contribution, $\Delta\phi$ due to a surface layer of electrical dipoles. Owing to the thermal agitation of the dipoles, the effective moment of the layer will vary, and by analogy with Langevin's theory of paramagnetism, $\Delta\phi$ may be shown to vary inversely as the absolute temperature, i.e. $\frac{e}{k}\ \Delta\phi = \frac{\gamma}{T}$. Hence

$$b^x = b_0 + \gamma/T$$

$$= b_0 + \frac{\gamma}{T_0} - \frac{\gamma}{T_0^2}(T - T_0) = b + \beta T$$

where $b = b_0 + \frac{2\gamma}{T_0}$ and $\beta = -\gamma/T_0^2$

Thus $\log_e A = \log_e A_0 + \gamma/T_0^2$

and $b = b_0 + 2T_0 \log_e A/A_0$

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in accordance with the experimental result. An important and necessary point in Zwikker's theory is that the dipoles are permanent dipoles in the atoms or molecules of the surface layer, and are not formed by ions and their induced images.

A discussion of Bridgman's theory of the relation between the deviations of A from the " universal " value, the temperature shift of the photo-electric threshold frequency and the surface heat of charging, is given by HERZFELD, *The Surface Heat of Charging* (1.540). He concludes that at least part of the shift in the threshold frequency is explained by the ordinary thermal expansion of the material and the dependency of the work function on the volume. Numerical calculations for the changes of A and the threshold frequency are compared with experiment. HERZFELD states that no complete explanation can be given until the space charge effect of the electrons in the transition layer at the surface of the metal can be calculated.

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SECTION 2

VARIATION WITH TEMPERATURE OF SPECIFIC ELECTRON EMISSION IN VACUO AND VALUES OF THE RICHARDSON CONSTANTS*

THE work carried out prior to the researches of LANGMUIR and others which disclosed the serious effects of residual gas and of space charge limitation of current on the measured emission, although successful in establishing the validity of a temperature variation in accordance with Richardson's formula, cannot be said to give reliable information as to the values of the constants in the formula corresponding to specific emission from the substances investigated.

LANGMUIR, *The Effect of Space Charge and Residual Gases on Thermionic Currents in High Vacuum* (2.100), lays down the con-

* Two main formulæ, both due in the first place to Richardson, have been used to represent the temperature variation of thermionic emission.

$$(1)\quad i = a\,T^{\frac{1}{2}}\,e^{-\frac{B}{T}}$$

(a and B constants.)

$$(2)\quad i = A\,T^{2}\,e^{-\frac{b}{T}}$$

(A and b constants.)

Over the temperature ranges worked with in practice, both formulæ represent equally well the experimental data. If the constants in one formula have been obtained by fitting on to a certain set of measurements, the constants for the other formula can be obtained immediately from the following equations:

$$B - b = \frac{3}{2}T_m$$

$$\log_{10} a - \log_{10} A = \frac{3}{2}\log_{10} T_m + \frac{3}{2\times 2\cdot 303}$$

where T_m is the mean temperature in the experimental range. Formula (2) is theoretically preferable to formula (1), and is used exclusively in the recent work.

The constants b and B have the dimensions of temperature and are expressed in degrees, a and A are usually expressed in amps./cm.^{2} deg.$^{\frac{1}{2}}$ and amps./cm.^{2} deg.^{2} respectively, so that the emission current i is obtained in amps./cm.^{2}

In some cases, instead of the constants b or B, use is made of electron potential energies $\phi = \frac{bk}{e}$ or $\Phi = \frac{Bk}{e}$. The numerical connection between the constants is as follows:

ϕ (or Φ) in volts $= 8\cdot 62 \times 10^{-5}\,b$ (or B) in degrees.

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ditions necessary to eliminate important secondary effects, in the following words :—

(1) Extremely high vacuum ; that is, a pressure below 0.001 mm., should be obtained. The presence of certain gases is much more injurious than others. Gases such as oxygen, water vapour, carbon dioxide, and hydrocarbons, which are very active chemically at high temperatures, should be especially avoided. This means that all stopcocks and sealing-wax must be eliminated and all glass parts, not to be cooled by liquid air, must be heated for at least an hour to 360° C. or more. Even with the Gaede molecular pump these precautions are necessary.

(2) Avoidance of large anodes, except those that have been especially treated by heating in a vacuum to 2,000° C. or have been exposed to powerful electron bombardment in a very high vacuum. Treating the metal by making it an electrode in an ordinary glow discharge except when the inert gases are used, is about the worst thing that could be done to it. Preferably, the anode should consist of tungsten wire which is freed from gas by heating to 2,500° C. for ten minutes. This should be done in the apparatus itself.

(3) The relative position and size of the electrodes should be such that space charge does not limit the current to an undesirable degree.

LANGMUIR'S experiments show that for tungsten in a perfect vacuum the emission current is given by

$$34 \times 10^6 \, T^{\frac{1}{2}} \exp \{ - 55{,}500/T \} \text{ amps./cm.}^2$$

In an appendix to his paper LANGMUIR gives a number of preliminary values of the thermionic constants for five substances.

$$A \, T^{\frac{1}{2}} \exp (- b/T) = i \text{ amps. /cm.}^2$$

Substance.	Thermionic currents at 2,000° K. milliamps./cm.².	b.
Tungsten	3	55,000
Tantalum	7	50,000
Molybdenum	13	50,000
Platinum	0·6	80,000
Carbon	1·0	32,000

The values of b are stated to be probably a little too high. LANGMUIR, *Thermionic Currents in High Vacua. I. Space Charge Effects. II. Emission of Tungsten and the Effect of Residual Gases* (2.101), a little later gives for tungsten the modified values

A	b
23.6×10^6	52,500

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K. K. SMITH, *Negative Thermionic Currents from Tungsten* (2.110), in 1915 published an account of measurements of the thermionic currents from tungsten filaments which had been subjected to various heat treatments. Somewhat different results were obtained, depending on the previous heat history of the filament, but the general conclusion was that under the best vacuum conditions Richardson's formula held good, the temperature range worked with being 1050° K. to 2,300° K. Values for A and b for tungsten were found as follows :—

$$\{A\ T^{\frac{1}{2}} \exp(-b/T)\} = i \text{ amps./cm.}^2$$
$$A = 67 \times 10^7$$
$$b = 54{,}700$$

A filament which had never been heated to 2,700° K, however, gave values :—

$$A = 44 \times 10^7$$
$$b = 59{,}700$$

SMITH points out that his experimental results are represented equally well using the fomula

$$i = C\ T^2 \exp(-d/T)$$

In SCHLICHTER'S work, *Spontaneous Electron Emission from Glowing Metals and the Thermionic Element* (2.120), on the emission from platinum and nickel, the spontaneous current (i.e. current in the absence of applied field) between the emitter and a cold electrode which practically enveloped it, was determined, in addition to the usual saturation current. An electric oven served to heat the emitter, and the collector electrode was water-cooled. For pure platinum, after the effects due to positive emission had decayed, the saturation and spontaneous currents were found to be in a constant ratio independent of temperature, and approximating to unity. For nickel the ratio of spontaneous to saturation current varied with temperature.

For the constants in the Richardson formula, saturation current

$$i = A\ T^{\frac{1}{2}} \exp(-b/T) \text{ amps./cm.}^2$$

SCHLICHTER found :—

	A.	b.	Temperature Range.
Platinum..	$1{\cdot}25 \times 10^7$	51,100	1,200–1,500° K.
Nickel	$7{\cdot}6 \times 10^6$	34,000	1,200–1.300° K.

RICHARDSON, *The Influence of Gases on the Emission of Electrons and Ions from Hot Metals* (2.125), drew attention to an interesting relation between pairs of values of A and b obtained by different

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investigators for various platinum and tungsten filaments. The conditions under which the different pairs of values were determined (the heat treatment of the filament, the presence or absence of gases) varied widely. In each case the relation $\log A = \alpha b + \beta$ held good, where α and β represent constants characteristic of the emitting substance (Pt or W), α being positive.

Further preliminary values of the Richardson constants based on results obtained by DUSHMAN are quoted by LANGMUIR, *The Relation between Contact Potentials and Electro-chemical Action* (2.130), in a paper which also contains a further discussion of the work of PRING and PARKER and others (0.202).

Substance.	A.	*b*.
Tungsten	$2{\cdot}36 \times 10^7$	52,500
Thorium	2×10^8	39,000
Tantalum	$1{\cdot}12 \times 10^7$	50,000
Molybdenum	1×10^7	50,000
Carbon (untreated)	—	48,000
Titanium	1,300 (?)	28,000 (?)
Iron	2,400 (?)	37,000 (?)

The emission from molybdenum under different gaseous conditions, and for wires subjected to different heat treatments, was examined by STOECKLE, *Thermionic Currents from Molybdenum.* (2.140), *Remarks on a paper by Mr. E. R. Stoeckle, entitled, "Thermionic Currents from Molybdenum"* (2.141), using a specially designed tube embodying water cooling of surfaces subjected to radiation from the filament, a precaution to minimise the liberation of adsorbed gases. The tube was kept on the pumps. In the final form of apparatus a pressure of 5.10^{-5} mm. could be held with a filament temperature of 2,300° K. A number of widely different pairs of values for the constants A and *b* under different vacuum conditions were obtained which, however, conformed to the relation $\log A - \alpha b + \beta$ where α and β denote constants. The most probable values for the specific emission appeared to be

$$A = 1.1 \times 10^8,$$
$$b = 53{,}600$$
$$\{i = A\, T^{\frac{1}{2}} \exp(-b/T) \text{ amps./cm.}^2\}$$

An extensive investigation by GERMERSHAUSEN, *Electron Emission of CaO Electrode in Vacuum* (2.150), *Electron Emission of Calcium Oxide in Gases and in High Vacuum* (2.151), of the emission from CaO in vacuum and in different gases, established conclusively that oxides show a pure temperature emission precisely similar to

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that for metals. The following values of the constants in Richardson's equation

$$i = A\,T^{\frac{1}{2}} \exp(-b/T)$$

are given by GERMERSHAUSEN for CaO.

$$A = 7 \times 10^6 \text{ amps/cm.}^2 \text{ deg}^{\frac{1}{2}}.$$

$$\frac{kb}{e} = 2.5 \text{ volts}$$

$$b = 29{,}500.$$

Commenting on the fact that earlier workers (WEHNELT, DEININGER, JENTZSCH, SCHNEIDER) found values of kb/e equal to 3.67, 3.76, 3.48, 3.45 volts respectively, GERMERSHAUSEN points out that these were obtained under imperfect vacuum conditions and that his own researches on the emission in the presence of hydrogen, lead to values of kb/e equal to 3.45 and 3.2 volts, the latter corresponding to measurements in hydrogen of a higher degree of purity.

A valuable critical review of the measurements of temperature variation made prior to 1916, is contained in RICHARDSON'S treatise, *On the Emission of Electricity from Hot Bodies* (2.160), the first edition of which appeared in 1916.

Measurements by HÜTTEMANN, *Emission of Electrons and Positive Ions by Glowing Filaments* (2.170), carried out before 1914, but delayed in publication until 1917, on the metals tungsten and tantalum lead to values of the constant b ($i = AT^{\frac{1}{2}} \exp(-b/T)$) equal to 57,100 and 49,100 respectively. Excellent confirmation of the Richardson equation over the range 1,000 to 2,000° C. was obtained.

An investigation published by STEAD, *The Short Tungsten Filament as a Source of Light and Electrons* (2.180), in 1920, was directed mainly to the determination of data for use in the design of thermionic valves employing the electronic emission of "pure" tungsten (i.e. ordinary commercial pure metal) or of 1 per cent. thoria tungsten. Only the results for the former are given, however. The principal interest was the determination of end corrections, but the measurements lead to the allocation of values for the constants in Richardson's formula. Stead does not actually evaluate these, but by plotting his results in a suitable form for deriving A and b, two straight lines are obtained which correspond to the Richardson Law with different pairs of values for the constants over a high temperature range (2,500—2,700° K.) and a low temperature range (2,000–2,500° K.).

$$i = AT^2 \exp(-b/T)$$

—	A (amps./cm.^{2}).	b.
Low temperature range	2·95	37,900
High temperature range	32·4 × 10^2	55,200

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The filament used for the emission measurements was a straight wire along the axis of a cylindrical anode fitted with guard ring extensions. The filament temperature was measured using a Cambridge Optical Pyrometer, and the absorption of the glass wall of the tube was allowed for. The emissive power of tungsten at different temperatures used in the calculation of the true temperature was that determined by WORTHING.

ARNOLD, *Phenomena in Oxide-Coated Filament Electron Tubes* (2.190), gives a review of the work carried out in the Western Electric Company's Laboratories on the development of oxide-coated filament valves for telephone repeater service. The average emission constants based on temperature variation measurement on 4,301 samples are given as

$$b = (19{\cdot}4 \text{ to } 23{\cdot}8) \times 10^3$$
$$\phi = kb/e = 1{\cdot}55 \text{ to } 1{\cdot}90 \text{ volts}$$
$$A = (8 \text{ to } 24) \times 10^4$$
$$i = AT^{\frac{1}{2}} \exp(-b/T) \text{ amps./cm.}^2$$

Measurements of the temperature variation of the emission from a tungsten filament were made by DAVISSON and GERMER, *The Thermionic Work Function of Tungsten* (2.200), in a research on the work function. The paper is summarised in detail elsewhere, but the values of the constant b obtained may with advantage be given here

$\{i = AT^{\frac{1}{2}} \exp(-b/T)\}$

	Temperature Scale.
$b = 55{,}410$	Worthing and Forsythe.
$b = 56{,}460$	Langmuir.

These values were obtained assuming a diameter of the filament $7{\cdot}553\ 10^{-3}$ cms. in place of the measured value $7{\cdot}595\ 10^{-3}$ as with the former, the temperature measurements become more consistent, and it was considered that loss by evaporation might reasonably be expected to account for the difference. No correction was made for the Schottky effect, but in a later note the authors remedy this defect, the revised value for b (Worthing and Forsythe) becoming 55,900.

DUSHMAN and EWALD, *Graphs for Calculation of Electron Emission from Tungsten—Thoriated Tungsten, Molybdenum and Tantalum* (2.210), in an explanatory paper dealing with the design of valves, quote values for the thermionic emission constants of a number of substances, based on results to be published later by DUSHMAN, ROWE and KIDNER. The b values given are:—

Thoriated tungsten	34,100
Tungsten	52,600
Molybdenum	50,000
Tantalum	46,500*
Thorium	34,100
Calcium	26,100

* First approximation.

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The emission at any temperature is calculated from the formula i in amps./cm.^{2} $= 60 \cdot 2\ T^2 \exp(-b/T)$. The results are based on the Forsythe-Worthing temperature scale for tungsten.

A preliminary abstract of Dushman, Rowe and Kidner's measurements is contained in a communication of 1923, the complete paper is discussed below, *The Electron Emission from Tungsten Thorium, Molybdenum and Tantalum* (2.211).

DAVISSON and GERMER, *The Thermionic Work Functions of Oxide-coated Platinum* (2.220), carried out for oxide-coated platinum a similar investigation to that already made by them for tungsten. They obtained for bk/e the value 1·79 volts, which agreed well with the work function deduced from calorimetric measurements (see 8.200).

The variation of the electronic emission of certain metals over a temperature range including the melting point of crystallographic transition points was investigated by GOETZ, *Electron Emission at the Melting Point and at Transition Points* (2.230, 2.231). GOETZ found that for the metals, copper, iron and manganese at the melting point, the constant b in the emission formula increased, although the emission itself exhibited no discontinuity. The constant A must, therefore, have varied as well and by just that amount necessary to counteract at the melting point the change in b. For the transformation of γ iron into δ iron, a variation of A only took place, and the temperature emission curve contained a discontinuity. Values of the constants are not given explicitly.

The difficult problem of the thermionic emission of the alkali metals was attacked by YOUNG, *The Thermionic and Photo-electric Properties of the Electro-positive Metals* (2.240), in 1924. He measured the electron current from potassium at temperatures up to about 250° C., and by employing a quartz tube was able to determine the photo-electric threshold frequency under the same experimental conditions. For temperatures above 125° C. the thermionic current was sufficiently large to permit of the use of a galvanometer, and working at the fixed anode potential of 370 volts a number of sets of observations were made, leading to values of the constants A and b, ranging from (2 to 3,100) $\times 10^{20}$ for A and (1·15 to 1·42) $\times 10^4$ for b ($i = AT^{\frac{1}{2}} \exp\{-b/T\}$). The mean values for the eleven series of observations gave $A = 480 \times 10^{20}$, $b = 12{,}900$. For temperatures less than 125° C. an electroscope was used to measure the current and a number of anomalous effects were obtained; in particular, the Richardson formula was found not to hold good. The chief point of interest in the results at lower temperatures is the continuity in the temperature variation of the emission in the neighbourhood of the melting point of potassium 62·5° C., no sharp change in the law being observed.

Further measurements of the thermionic and photo-electric emission from potassium are described in a paper by RICHARDSON and YOUNG, *The Thermionic Work-Functions and Photo-electric*

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Thresholds of the Alkali Metals 2.241), which is directed principally to the question of the existence of multiple thresholds for sodium and potassium. A theory is advanced to account for the different behaviour of potassium in thermionic and photo-electric measurements, depending on the idea that the metal surface is divided into patches having different work functions. If the total area of the patches having the lower work function is very small compared with the total area of the surface, the photo-electric currents will be determined almost entirely by the emission from the higher work function regions. The thermionic effect acts differently, however, and the lower work function patches will dominate the emission at lower temperatures. The effect of hydrogen and water vapour on the emission from potassium was determined and the results fitted in with the patch theory. It was observed that the values of the thermionic constants A and b obtained under different conditions satisfy approximately a linear relation between log A and b.

Values of the Richardson constants for a large group of metallic oxides were obtained by SPANNER, *Thermal Emission of Electrically-charged Particles* (2.250). (See 8.300 for a summary of Spanner's results.)

In 1925, accounts of two researches were published, one by ZWIKKER and the other by DUSHMAN, ROWE, EWALD and KIDNER, which may be regarded as giving the most careful determinations of the constants in the Richardson formula yet available. A fairly detailed account will therefore be given. We consider first the work of DUSHMAN and his collaborators, *The Electron Emission from Tungsten, Molybdenum and Tantalum* (2.260).

In the revised Richardson formula $i = AT^2 \exp(-b/T)$, theoretical considerations lead to the suggestion that A is a universal constant, at least for elementary substances. Its value derived on somewhat different assumptions is given by one or the other of the formulæ

$$A = \frac{2\pi k^2 me}{h^3} = 60 \cdot 2 \text{ (Sackur-Tetrode).}$$

$$A = \frac{2^{5/2}\pi^{9/2}\varepsilon^{-5/2}}{15} \frac{k^2 me}{h^3} = 51 \cdot 2 \text{ (Lewis, Gibson and Latimer).}$$

ε = base of natural logarithms = 2·72.

It was one object of Dushman's work to differentiate, if possible, between these two forms, and a very high degree of accuracy was aimed at.

The elements investigated were tungsten, molybdenum and tantalum. A V-shaped filament of the metal was mounted in a lime-glass tube, the anode being formed by depositing from a subsidiary calcium-covered tungsten filament a film of metallic calcium on the interior wall of the tube. Before sealing off, the tube was baked at 360° with the pumps on, for an hour, allowed to

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cool, and the filament then flashed at a high temperature (about 2,800° K. for tungsten), and run for an hour at a somewhat lower temperature until all gas evolution had ceased. After sealing the tube and ageing the filament for 15–24 hours, the tube was immersed in liquid air and the filament flashed at a high temperature for one minute. The flashing at high temperature was usually repeated in between emission current measurements made at different temperatures.

For determining the temperature of the filament in the thermionic tube the following method was employed. Special vacuum lamps were made by using V-shaped filaments of the same kind of wire as in the actual thermionic tube, and the current-temperature relation for these filaments was then obtained by pyrometering against a calibrated standard tungsten lamp furnished by WORTHING and FORSYTHE. For molybdenum and tantalum a knowledge of the emissivity compared with that of tungsten is involved in the calculation. For tantalum, Worthing's emissivity results were employed; the data of FOOTE and FAIRCHILD would have led to temperatures differing a little from those actually taken (discrepancy at 2,000° K. = 24°). The current-temperature relations found in this way were assumed to hold good for the filaments of the actual tubes.

To obtain from the measured emission current at a given temperature the true emission per unit area at that temperature, allowance must be made for the fact that the ends of the filament are at a lower temperature than the middle. The end correction formula used was that of WORTHING.

The SCHOTTKY effect (0.216) was taken into account by applying Schottky's formula :—

$$i_v = i_o \left(\frac{dV}{dx}\right)^{\frac{1}{2}} \exp\,(4.39/T),$$

i_v = current at anode voltage V

i_o = current at anode voltage O

$\frac{dV}{dx}$ = potential gradient at cathode.

Measurements made at different anode voltages confirmed a formula of this type and the value of i_o was found by extrapolating to V = o.

Taking logarithms of both sides of Richardon's revised equation we have :

$$\log_{10}\frac{i}{T^2} = \log_{10} A - (b/2.303T).$$

If A is put equal to 60.2, for each pair of values of i and T a value of b is obtained. Constancy of the b values so derived, for different temperatures, indicates the validity of Richardson's equation with A = 60.2. Alternatively plotting $\log_{10} i/T^2$ against

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1/2.303T the slope of the best straight line through the experimental points, gives a further determination of b. For tungsten filaments the values of b found by these methods were in excellent agreement, the best representative value being given as 52,600 $\pm$ 250. For tantalum, too, the results were in good agreement with an A value of 60.2, leading to b = 47,800 $\pm$ 500. For molybdenum, however, the values of b deduced by the first method showed a systematic decrease with increasing temperature, the average value being 50,230 against 52,770, deduced from the slope. The discrepancy is attributed by the authors to the presence of adsorbed oxygen on the surface of the filament, and it is claimed that the results cannot be regarded as evidence for or against the assignment to A of the value 60.2. The results for the first three tungsten wires were the most accurate and for these, any departure from the Richardson law with A = 60.2 lay within the experimental temperature error.

The following table summarises the results obtained :—

Metal.	Mean Value of b by Equation with A = 60·2.	Least Square Method.	
		b.	Log_{10} A.
Tungsten. Specially-prepared pure. Sample I	52,580	52,560	1·773
Tungsten. Specially-prepared pure. Sample II	52,660	52,570	1·753
Unthoriated Tungsten, as used in regular lamp production	52,830	52,860	1·787
Pure Tungsten. Sample I. Using a modified tube employing a guard ring anode to avoid end corrections.	52,470	51,390	1·532
Heat-treated tungsten containing very long crystals	52,680	53,850	2·034
Average	52,640	52,650	1·776
Molybdenum—			
" Regular " wire	50,230	52,770	2·404
" Regular " wire	49,730	51,180	2·133
Average	49,980	51,980	2·268
Tantalum (Osram Co., Germany) ..	47,700	46,840	1·567
Tantalum (Fansteel Co., Chicago) ..	48,000	47,550	1·663
Average	47,850	47,200	1·615

ZWIKKER, *Physical Properties of Tungsten at High Temperatures* (2.270), using a specially designed system of bulbs, made a comparison of the effects on the emission current of a given tungsten filament, run at constant current and constant anode voltage, of the various

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methods used to obtain good high vacuum conditions. Cleaning up by evaporation of a subsidiary tungsten filament was found to be as effective as vaporising magnesium in the tube, and both were better than connecting the tube with a bulb containing charcoal plunged in liquid air. ZWIKKER discusses a number of other points dealing with the effect on the emission of different running conditions.

For the main measurements, the temperatures were determined by the wattage consumption of filament which is known as a function of temperature for a non-emitting filament. When the filament is emitting, input and output currents are different and a suitable mean must be taken. In addition, the energy used up in transferring the electrons from inside to outside, gives a lower temperature for the same watts consumption. Both these effects were allowed for in determining the filament temperature. The small correction for the heating of the filament by radiation from the anode (made hot by electronic bombardment) was measured and found to be negligible at temperatures less than 2,400 ; for higher temperatures, however, it was taken into account. A correction to the emission current was made to allow for the Schottky effect. Two tubes with emitter wire 99.17 μ, and three with emitter wire 193.3 μ, were worked with. The mean results for the whole series are given in the following table :—

T.	Emission per unit area for zero anode voltage.
2,500° K.	0·324 amps./cm.^{2}
2,400	0·119
2,300	0·0425
2,200	0·0140
2,100	0·00398
2,000	0·000978
1,900	0·000211
1,800	0·000348
1,700	0·00000433

These values are represented by the formula

$$i = 60{\cdot}2\ T^2 \exp(-52{,}230/T) \text{ amps./cm.}^2$$

The data do not, however, differentiate between the two theoretical values of A, 60.2 and 51.2.

$$\Phi = \frac{bk}{e} = 52{,}230 \times 8.62 \times 10^{-5} = 4.50 \text{ volts.}$$

These results lie intermediate between those of DAVISSON and GERMER and those of DUSHMAN and EWALD. Reasons are given for thinking that DAVISSON and GERMER took the diameter of their

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wire too large, which would give too low temperatures. Dushman and Ewald, and Zwikker's curves correspond to a temperature difference of 10° at 2,300° and 0° at 1,900°.

ZWIKKER, *Thermionic Emission of the Metals Tungsten, Molybdenum, Thorium, Zirconium and Hafnium* (2.271) also made measurements on the metals molybdenum, thorium, zirconium and hafnium.

The values of the constants A and b in the formula

$$i = AT^2 \exp(-b/T)$$

which fitted best the experimental data, are given in the following table :—

Metal.	A.	b.
Tungsten	60·2	52,225
Molybdenum	60·2	50,900
Thorium	60·2	38,900
Zirconium	330·	47,900
Hafnium	14·5	41,000

HARRISON, *A Study of the Concurrent Variations in the Thermionic and Photo-electric Emission from Platinum and Tungsten with the State of the Surface of these Metals* (2.280), working on the emission from the metals platinum and tungsten, obtained values for the work functions which depended on the previous history of the filament. In particular, for platinum two types of emission were distinguished depending on whether the platinum was in a " large-emission " or a " small-emission " state. Values of bk/e for tungsten were obtained ranging from 4.78 to 6.57 volts. For platinum in the large-emission state, bk/e varied in different sets of measurements on a given strip from 3.42 to 4.07 volts, and in the small-emission state from 3.52 to 5.17 for the same strip.

WARNER, *A Comparison of the Thermionic and Photo-electric Work Functions for Clean Tungsten* (2.281), measured the variation of thermionic emission with temperature in order to compare with the photo-electric work function of the same filament measured at a temperature at which the thermionic current had practically ceased. For bk/e Warner obtained 4.71 volts by thermionic experiments, against 4.79 volts deduced from his photo-electric measurements.

The use of the formula $i = A \exp(-b/T)$ to represent temperature emission data is urged by HAM, *Interpretation of Data dealing with Thermionic Emission* (2.289), who found Davisson and Germer's data satisfactorily represented in this way.

GOETZ, *Electron Emission of Metals during Change of State* (2.290, 2.291), describes further measurements of the emission of metals in a temperature range, including the melting point. A particularly

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full and detailed description is given of the experimental arrangements, including the construction of a high vacuum melting oven for temperatures up to 2,000° C. GOETZ found for silver, gold and copper that, in passing through the melting point to higher temperatures, the work function dropped discontinuously and then rose gradually to a value which for Ag and Au equalled that for the solid state and for copper equalled about 4/3 of the solid state value. In the liquid state both " constants " b and A depended on temperature. For the solid crystalline states of the metals the constant A ($i = AT^2 \exp(-b/T)$) agreed tolerably well with the theoretical value 60·2 (in ampere c.g.s. units).

Some further evidence on the effect of change of state on thermionic emission was obtained by CARDWELL, *The Photo-electric and Thermionic Properties of Iron* (2.292), who measured the emission from outgassed iron in passing through crystallographic transition points.

The emission constant b for the oxides of barium, calcium and strontium was determined by ESPE, *Exit Work of Alkaline Earth Oxides Cathodes* (2.300), by the temperature variation method. The following values were found for b :—

CaO	22,400 ± 300
SrO	16,600 ± 250
BaO	12,900 ± 250

$$i = AT^{\frac{1}{2}} \exp(-b/T).$$

A relation between the values of ϕ the exit work, for emitters in different groups of the periodic table, was derived by MICHEL, *Exit Work of Thermionic Electrons* (2.305). Michel's result runs $\phi\delta^2 =$ constant, where δ is the distance between two adjacent atoms in the surface of the emitter, and the constant has a value characteristic of the group. Comparison with Spanner's and Michel and Spanner's measurements indicates good agreement with the formula for the second group and tolerable agreement for the third and fourth groups.

Michel's result was criticised by SMEKAL, *Exit Work of Thermionic Electrons* (2.306), who points out that Espe's experimental data accord better with the formula $b\delta =$ constant. The use of the oxide lattice rather than the metal lattice, in determining δ, is also criticised. MICHEL (2.307) replied to these criticisms.

The thermionic properties of platinum had always presented a difficult problem owing to the enormous effects of small traces of gas on the emission. DU BRIDGE, *The Photo-electric and Thermionic Work Function of Out-gassed Platinum* (2.310), subjected platinum filaments to a careful outgassing process, and followed the emission during the outgassing. He found consistent results obtainable only when the experimental tube was sealed off and gettered with magnesium. When this was done, reproducible values of the emission

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were obtained, and it was verified by observations on the same filament that the photo-electric and thermionic work functions were equal, a value of 6·35 volts being obtained. As regards the constant A in the emission equation $\{i = AT^2 \exp(-b/T)\}$, DU BRIDGE found that during outgassing log A was proportional to b, and that for the final outgassed state, A = 14,000 (amps./cm. deg.^{2}). This is some 200 times the theoretical value 60 given by DUSHMAN. DU BRIDGE, *Thermionic and Photo-electric Emission from Pt and Pd* (2.311), also measured the emission from palladium and found for the thermionic and photo-electric work functions, the common value 5.35 volts. This value was enormously decreased by slight traces of hydrogen.

DU BRIDGE, *The Thermionic Emission from Clean Platinum* (2.312), in order to establish definitely his work on platinum, repeated the thermionic emission determinations over a more extended temperature range (1,360–1,750° K.) and improved the accuracy of measurement. He found, in general confirmation of his earlier results, the following values for the thermionic constants :

$$A = 17{,}000\,;\ b = 72{,}800\,;\ \phi = 6{\cdot}27 \text{ volts.}$$

A further empirical relation between the thermionic emission constants was pointed out by DU BRIDGE, *Systematic Variation of the Constant A in Thermionic Emission* (2.315). The relation is an extension of that discovered by RICHARDSON in 1915. DU BRIDGE found that the straight lines $\log A = \alpha b + \beta$ corresponding to concurrent variations of the thermionic constants A and b of a given substance, all intersect each other in the neighbourhood of the point A = 60, b = 52,000. This holds for platinum, tungsten, thoriated tungsten and liquid copper, potassium, however, being an exception. The theoretical consequences of the relation are indicated by DU BRIDGE.

CARDWELL, *Effect of a Crystallographic Transformation on the Photo-electric and Thermionic Emission from Cobalt* (2.320), found a marked change in photo-electric and thermionic properties of cobalt at a temperature of approximately 850° C., and concluded that the work function for cubic cobalt is less than that for the hexagonal form.

In experiments on the photo-electric and thermionic properties of metals which had been subjected to rigorous heat treatment under extreme vacuum conditions, MARTIN, *The Photo-electric and Thermionic Properties of Molybdenum* (2.330), obtained values for the thermionic and photo-electric work functions of molybdenum. For the thermionic effect the slope of the Richardson line gave $\phi = 3{\cdot}48 \pm 0{\cdot}07$ volts. The average value of the photo-electric work function was found to equal $3{\cdot}22 \pm 0{\cdot}16$ volts. The two work functions are sufficiently near to be taken as identical, having regard to the experimental errors of the measurements.

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Investigating the thermionic and photo-electric properties o rhodium during and after subjection to rigorous heat treatment and the effect of gases on the photo-electric emission, DIXON, *Som Photo-electric and Thermionic Properties of Rhodium* (2.332), found a thermionic work function ϕ equal to $4{\cdot}58 \pm 0{\cdot}09$ volts. This resul agrees with the photo-electric work function at 240° C., $4{\cdot}57 \pm 0{\cdot}0$ the value at 25° C. being $4{\cdot}92 \pm 0{\cdot}06$ volts. A noticeable featur of the thermionic measurements was an irregularity in the emissio current at about 1,100° C., indicating a structure change in rhodiun at this temperature with liberation of heat on passing to lowe temperatures.

An excellent collection of data on thermionic emission is con tained in a contribution by DUSHMAN to the *International Critica Table* (2.400).

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SECTION 3

HEAT EFFECTS IN THERMIONIC EMISSION

IT was pointed out by RICHARDSON in his fundamental memoir, that in the passing out of electrons from the surface of a hot body a certain amount of energy is lost. The first experimental investigations of this effect, which is manifested as a cooling of the emitter, are due to JENTZSCH, and WEHNELT and JENTZSCH. Their work and that of RICHARDSON, COOKE, SCHNEIDER, etc., carried out prior to 1915, is summarised in Section O (0.180–0.188).

The first investigation of the cooling effect, in which the conditions necessary to obtain reliable thermionic measurements as indicated by LANGMUIR were observed, is due to LESTER, *The Determination of the Work Function when an Electron escapes from the Surface of a Hot Body* (3.100). LESTER gives a valuable critical discussion of the earlier work of RICHARDSON and COOKE. In his experiments the bridge method was used, in which the thermionic emitter filament forms one arm of a Wheatstone's bridge. The same method had previously been adopted by RICHARDSON and COOKE.

The change of resistance due to the cooling effect was measured and compared with that produced by a known change of electrical energy. The theory of the method and certain necessary corrections are contained in the paper. LESTER sets the measured cooling effect equal to $\phi + \frac{2k}{e}(T - T_0)$ in equivalent volts and concludes from a comparison of his results for tungsten, with the coefficient b in Richardson's equation $i = AT^{\frac{1}{2}} \exp\left(-\frac{b}{T}\right)$, as measured by LANGMUIR, that $\frac{bk}{e} = \phi$. For tungsten, ϕ by the cooling effect equals 4·478 volts (mean of eighteen determinations), which corresponds to $b = 52{,}130$ against 52,500 given by LANGMUIR on the basis of temperature variation measurements. The term $\frac{2k}{e}(T - T_0)$, where T and T_0 are equal, respectively, to the temperature of the hot and cold electrodes, amounted to about 0·30 in equivalent volts. (In accordance with later theories (see RICHARDSON, 1.220) the expression for the cooling effect must be modified.)

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Lester's results are summarised in the following table :—

Emitter.	ϕ (volts).	$b = e\phi/k$.
Carbon	4·55	52,800
Molybdenum	4·59	53,200
Tantalum	4·51	52,500
Tungsten	4·478	52,130

The effects of traces of active gases, on the value of ϕ from the cooling effect, were found to be similar to the effects on b due to the same cause, discussed by LANGMUIR.

The results of measurements of the cooling effect for six coated filaments are reported by WILSON, *The Loss of Energy of Wehnelt Cathodes by Electron Emission* (3.110), who also determined the coefficient b by the temperature variation method. The values found by WILSON are given in the following table :—

Filament.	bk/e $\{i = AT^{\frac{1}{2}} \exp(-b/T)\}$.	ϕ.
1	2·34	2·39
2	2·59	2·54
3	2·02	1·97
4	2·16	2·28
5	3·28	3·22
6	3·49	3·45

1 and 2—BaO 50 per cent., SrO 25 per cent., CaO 25 per cent.
3 and 4—BaO 50 per cent., SrO 50 per cent.
5 and 6—CaO 100 per cent.

WILSON points out that his results establish the practical identity of ϕ as deduced from calorimetric and temperature variation measurements.

The most accurate determinations of the cooling effect due to thermionic emission at present available are those of DAVISSON and GERMER, *The Thermionic Work Function of Tungsten* (3.120). These workers used the same filament for determinations of ϕ by the calorimetric method and by the temperature variation method. In the calorimetric determinations an elaboration of the bridge method (see LESTER, 3.100) was employed. The change in voltage ΔE across the ends of a segment of the filament was determined, such that the filament current remained unchanged when the anode, from being negative to the filament, was made positive, i.e. when the space current was allowed to pass.

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In that case

$$\phi = \left(\frac{2EI}{E - I\frac{dE}{dI}} \right) \frac{\Delta E}{i}$$

E = voltage across test segment.
I = current in filament.
i = electron space current.

on certain simplifying assumptions, namely :—

(1) That the resistance of the filament (i.e. E/I) is dependent only on temperature, and not on the passage or otherwise of space current ;

(2) That the space current is so small that it is legitimate to treat the filament current as constant along the filament ;

(3) That all the voltage change ΔE arises from the cooling effect following the switching on of the space current.

The assumption (3) is, of course, not valid. Other effects which lead to a voltage change are divided into those which reverse in sign on reversing the direction of the heating current and those for which this is not the case. The former are eliminated by taking observations with the heating current in each direction in turn.

Of the irreversible effects may be mentioned those arising from (*a*) heating up of plates and filament due to electron bombardment, (*b*) asymmetric emission from the filament due to the asymmetry of the potential field caused by filament potential drop and the guard box potentials.

(*a*) is allowed for by determining the value of $\Delta E/i$ at constant filament current I for different plate voltages (to which the effect is proportional). The true value is then found by extrapolation to $V = 0$.

(*b*) This could have been avoided by making the guard boxes negative by the same amount to the respective ends of the filament instead of maintaining them at the same potential. The effect was, however, allowed for by finding the difference in emission for these two cases.

A further correction is necessary on account of the failure of assumption (2). The correction is examined in detail and it appears that it depends fundamentally on where the space current is led back into the filament. If this is done at the midpoint the voltage charge arising from this cause is

$$\Delta E^1 = +\frac{\beta}{14} \rho^1 L i^2.$$

If, on the other hand, the current is returned to the ends of the filament, $\Delta E^1 = -\frac{\beta}{12} \rho^1 L i^2$. As this is of different sign from the previous expression, it is possible by controlling the proportions of the current which return respectively to the centre and to the ends of the filament to contrive that the spurious voltage change shall be zero.

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The characteristics of the thermionic tube and circuit arrangements may be summarised as follows :—Filament : straight wire of high purity tungsten kept taut by molybdenum springs at the ends. Anode : two plates of molybdenum ; guard boxes surrounded the filament ends and delimited the emission. Tube : prepared by elaborate bakings and glowings of metal parts *in vacuo* ; finally sealed with molybdenum plates white hot, at a pressure of $1 \cdot 10^{-6}$ mm. Final pressure as measured with Buckley ionisation manometer, $5 \cdot 10^{-7}$ mm. (filament hot) and $5 \cdot 10^{-8}$ mm. (filament cold). Anode voltage = 196.

The final corrected value of $\phi = bk/e$ at 2,270° K. is given as 4·91 volts, the most unfavourable cumulative error being estimated as $\pm$ 0·045 volts, i.e. about 1 per cent.

The measurements of b ($i = AT^{\frac{1}{2}} \exp(-b/T)$) by the temperature variation method were made, using the same filament as for the calorimetric work. Again, an anode voltage of 196 was employed. The result obtained depends on whether the Langmuir or Worthing and Forsythe temperature scale for tungsten be adopted. The values given are

——	b.	$\phi = kb/e$.
Langmuir Scale	$5{,}656 \times 10^4$	4·869
Worthing and Forsythe Scale ..	$5{,}541 \times 10^4$	4·778

The more reliable of these values is that using the Worthing and Forsythe scale, and for this the discrepancy between calorimetric and temperature variation determinations of amounts to 2·7 per cent. DAVISSON and GERMER conclude their paper with a discussion of the theoretical relation between the two values and regard their experimental results as indicating that in the metal the electrons possess little or no energy.

A similar investigation to that described above for tungsten was carried out by DAVISSON and GERMER for oxide-coated platinum, *The Thermionic Work Functions of Oxides-coated Platinum* (3·121).

Cooke and Richardson's method was used by MICHEL and SPANNER, *The Cooling Effect for Oxide Cathodes* (3.122), to measure the cooling effect for glowing oxide cathodes. The computed exit works for CaO, SrO, and BaO agreed satisfactorily with values derived by Spanner using the temperature variation method.

The exit work for a number of technical oxide cathode tubes was determined by ROTHE, *Exit Work for Oxide Cathodes* (3.123), using both temperature variation and cooling effect methods. The cooling effect value agreed approximately with the temperature variation value, provided the saturated current was used. For smaller currents the cooling effect found was anomalously greater.

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ROTHE also studied the change in the emission over long periods and concluded that the high emission of oxide cathodes is due to metallic particles produced by chemical decomposition, and deposited in the interstices of the oxide coating.

The correction for the Schottky effect was omitted by DAVISSON and GERMER, *A Note on the Thermionic Work Function of Tungsten* (3.124), in their investigation of the work function of tungsten. The necessary amendment is made by these workers in a note published in 1927. The value of ϕ by the temperature variation method now becomes 4·818 instead of 4·778 volts, the value by the cooling effect method (4·91) being, of course, unchanged. A further discussion of the theoretical relations of the two ϕ's is included.

A new method for determining heats of condensation of electrons and ions was elaborated by VAN VOORHIS and COMPTON, *Heats of Condensation of Electrons and Positive Ions on Molybdenum in Gas Discharges* (3.130, 3.131), who measured the rate of heating up of a small sphere of molybdenum in a region of intense gas ionisation, using a thermocouple. The space potential and mean electronic energies were found by using the sphere as an exploring electrode. The heat of electron condensation on molybdenum was found to be 4·76 volts in argon, 4·77 or 5·01 volts in nitrogen, 4·04 or 4·35 volts in hydrogen. The different values follow different treatments of the molybdenum surface.

ROTHE, *A New Circuit for the Measurement of the Cooling Effect* (3.140), describes a new circuit for the measurement of the cooling effect with greater accuracy than heretofore, and offers an explanation of the anomalously large cooling effect obtained by him in his previous work. The possibilities of measurements of entrance heat using oxide cathodes are discussed.

VIOHL, *Measurement of the Heat Developed when Electrons Condense on to Metals* (3.150), describes a method for the determination of the entrance heat or heat of condensation of electrons in a metal, and gives the results of measurements on nickel. The electrons emitted by an oxide cathode were accelerated by means of a magnetic field, and diaphragms were arranged in such a way that only electrons of a certain velocity could reach the receiver, which consisted of a metal block of low thermal capacity, the temperature of which could be determined by means of a copper-constantan thermo-element. The kinetic energy acquired in the field was evaluated from a determination of the velocity distribution of the electrons. This also enabled the contact potential difference to be allowed for. For nickel the condensation heat was found to correspond to 4·31 volts or 97,800 calories per mol.

An investigation bearing on the thermal effects accompanying thermionic emission was carried out by TIERI and RICCA, *Electronic Emission in a Vacuum Tube* (3.160). These workers used a triode valve and measured the changes in filament current ΔI (accuracy

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of one part in a million) and the changes in emission current Δi due to changing the anode potential from -110 to a given positive voltage. A series of pairs of values of ΔI and Δi for each anode potential was obtained by varying the grid potential. Special measurements at very low anode voltages showed a change in the sign of ΔI, ΔI equalling zero for an anode potential change from -110 to $+1{\cdot}4$ volts. The authors postpone the interpretation of their results.

VAN VOORHIS and COMPTON, *Heats of Condensation of Electrons on Several Metals in Several Ionised Gases* (3.169, 3.170), in a further investigation on the lines of their previous work (3.130, found that the heats of condensation of electrons on electrodes of molybdenum, platinum and tungsten coated with potassium, depended on the surrounding ionised gas even when this was a highly purified inert gas. The new results obtained are shown in the following table :—

Collector Material.	Argon.	Neon.	Helium.	Nitrogen.	Hydrogen + inert gas.
Mo	4·14	3·69	3·53	4·47	—
Pt	4·89	4·61	4·39	5·21	3·9
W + K	1·11	1·05	0·93	—	—

It was found to be very necessary to correct for changes in contact potential differences, due to the cleaning of the collector electrode by positive ion bombardment. This was particularly the case for platinum.

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SECTION 4

THE DISTRIBUTION OF VELOCITIES OF THERMIONIC ELECTRONS

It is to be expected from considerations of simple kinetic theory that the velocities of emission of thermionic electrons will be distributed in accordance with Maxwell's Law. This law states that if $N_u du$ be the number of electrons emitted having the velocity component normal to the emitting surface, lying between u and $u + du$, then

$$N_u\, du = N2hmu \exp(-hmu^2)\, du$$

where $h = 1/2kT$

and N = the total number of electrons emitted per sq. cm. per second,

and if any pair of rectangular axes in the surface are taken, then the number emitted with components along these axes lying between v and $v + dv$, and w and $w + dw$, are respectively

$$N_v dv = N\sqrt{\frac{hm}{\pi}} \exp(-hmv^2)\, dv,$$

$$N_w dw = N\sqrt{\frac{hm}{\pi}} \exp(-hmw^2)\, dw.$$

The first experiments made on the actual distribution of velocities among thermionic electrons were by Richardson and Brown, *The Kinetic Energy of the Negative Electrons Emitted by Hot Bodies* (4.100), in 1908, who measured the distribution as regards the normal component. Briefly, the method consisted in measuring the current between an emitting plate and an opposite parallel plate with various retarding potentials imposed across the condenser so formed. Treating the plates as infinite planes, Maxwell's Law would lead to a linear relation between the logarithm of the current and the retarding potential, the constants in the relation being determined by the temperature of the emitting surface and the current at zero retarding potential. Maxwell's Law was verified for platinum foil *in vacuo*. Tests with oxide-coated and hydrogen-saturated platinum gave distributions not in accordance with Maxwell's Law. Richardson, *The Kinetic Energy of the Ions Emitted by Hot Bodies* (4.101), *The Kinetic Energy of the Ions Emitted by Hot Bodies* (4.102), also tested the law as regards the velocity components parallel to the surface. He used an emitting strip parallel to a plate containing a slit. This plate could be moved as a whole

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in its own plane, so that the slit was displaced with respect to the emitting strip. These measurements, in conjunction with the results for the normal component, established the Maxwellian Law for the tangential components for platinum *in vacuo*.

The theory of the current between emitter and collector electrodes of different shapes, taking into account the Maxwellian distribution of the initial velocities, was worked out by RICHARDSON, ***On Thermionics*** (4.110).

In 1914, SCHOTTKY, *Emission of Electrons from Glowing Filaments under Retarding Potentials* (4.120), *Boundary Potentials for Cylindrical Electrodes* (4.121), investigated the law of distribution for electrons emitted by hot wires of tungsten and carbon. The heated wire formed the axis of a metal cylinder on which the emitted electrons were collected. SCHOTTKY, assuming Maxwell's Law, deduced the relation connecting the current between wire and cylinder, with the retarding potential across these two electrodes. The formula is applicable only at higher values of the retarding potential for which space charge effects are negligible. The critical potential at which such effects begin to occur was also deduced. In Schottky's experiments a special commutator, first introduced by O. V. BAEYER, *Slow Cathode Rays* (4.122), was used to cut out alternately heating and emission currents, so that the emission measurements were made only in the absence of the magnetic field and potential drop due to the filament current. The results on the whole confirmed Maxwell's Law for the emission velocity distribution. Certain deviations, however, were in the direction of an excess of high-velocity electrons.

A general discussion of the velocity distribution of electrons emitted in different ways—photo-electric effect, thermionic emission, etc., was given by BECKER, *Determination of Electron Exit Velocities* (4.130), *Comparison of Photo-electric and Thermionic Electron Emission* (4.131). He also goes into the theory of measurement of electron-velocities and derives the formulæ appropriate to different methods.

The use of a magnetic field parallel to the filament, to control the current between filament and anode, was discussed by HULL, *The Effect of a Uniform Magnetic Field on the Motion of Electrons between Coaxial Cylinders* (4.140), who showed that from the variation in the current with change in the strength of the magnetic field, any suggested distribution law could be tested. Preliminary measurements, using a tungsten filament *in vacuo*, were in good agreement with a Maxwellian distribution.

TING, *Experiments on Electron Emission from Hot Bodies* (4.150), used both the cylinder method of SCHOTTKY and also the parallel plate method, to determine the distribution law for electrons emitted by tungsten and platinum. Although the distribution found was Maxwellian, it corresponded to a temperature about double that of the emitting filament. JONES, *The Kinetic Energy*

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of Electrons Emitted from Hot Tungsten (4.151), continued Ting's research, confining his attention to tungsten. He failed to obtain the high values of the mean electron energy found by TING and attributed Ting's results to a possible effect of the mutual repulsion of the electrons (space charge effect) and to surface contamination.

A method for determining the "apparent" distribution of velocities of the emitted electrons by a study of the free potential of the grid of a triode valve is described by NUKIYAMA and KUWASHIMA, *The Free Potential of the Grid and the Plate of the Triode Valve and the Apparent Distribution of the Initial Electrons* (4.160).

POTTER, *The Distribution of Velocities among the Electrons Emitted by Hot Platinum in an Atmosphere of Hydrogen* (4.170), investigated the validity of Maxwell's distribution law for the electrons emitted by platinum both *in vacuo* and in hydrogen at pressures of 1/20 mm. and upwards. He found that the pressure of hydrogen disturbed the distribution law only in this way, that the mean energy of the emitted electrons was higher than that corresponding to the temperature of the platinum. A similar result to Potter's was obtained by CONGDON, *The Kinetic Energy of Electrons Emitted from a Hot Tungsten Filament in an Atmosphere of* (*a*) *Argon*, (*b*) *Hydrogen* (4.171), for tungsten in hydrogen. For tungsten in argon, however, both the form of the distribution law and the mean kinetic energy of the electrons were found to be in accordance with Maxwell's Law.

A careful test of the distribution law for tungsten was carried out by GERMER, *The Initial Energy of Thermionic Electrons* (4.180), *The Distribution of Inital Velocities among Thermionic Electrons* (4.181), using a modification of Schottky's method. Between the filament and the anode cylinder a coaxial cylindrical grid was inserted. By the application of suitable potentials to the electrode the effects of photo-electrons, or reflected electrons, were eliminated. Good verification of Maxwell's Law for tungsten *in vacuo* was obtained. In none of the eight distribution curves given could a better fit be arrived at by assuming in the formula a temperature differing by not less than 10 per cent. from the actual temperature of the wire.

Certain limitations to the applicability of Schottky's formula for the current between a straight filament and a coaxial cylindrical anode were pointed out by DAVISSON, *A Note on Schottky's Method of Determining the Distribution of Velocities among Thermionic Electrons* (4.189), who showed the space charge effect to be operative at higher values of the retarding potential than previously assumed.

The initial velocities of the electrons emitted thermally by alkaline earths were investigated by ROSSIGER, *The Velocity of Thermions from Alkaline Earths* (4.190). He used the following new method: the number of electrons is determined which pass through two staggered slits in two coaxial cylinders, the common axis being occupied by the emitting filament. A magnetic field

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parallel to the axis and a radial electric field act simultaneously on the electrons emitted. The controllable variables are the stagger angle, the magnetic field strength, and the electric field strength. From the variation in the current through the slits with each of these variables, the distribution law can be deduced. A Maxwellian distribution was obtained for the electrons emitted by a mixture of CaO, BaO and SrO.

KOLLER, *Electron Emission from Oxide-coated Filaments* (4.191), in an investigation of different aspects of the emission from barium and strontium carbonates (on a platinum iridium core) found Maxwell's Law obeyed as regards shape, but the average kinetic energy of the emitted electrons came out to be nearly 30 per cent. higher than that corresponding to the temperature of the filament. For oxide-coated platinum filaments, GERMER, *The Distribution of Velocities among Thermionic Electrons* (4.192), states that he found Maxwell's Law obeyed in both respects.

The question of the effect of gases on the velocity distribution law was taken up again by DEL ROSARIO, *The Effect of a Hydrogen Atmosphere on the Velocity Distribution among Thermionic Electrons* (4.200), who found, for tungsten and platinum in hydrogen up to 0·25 mm. pressure, complete obedience to Maxwell's Law, the hydrogen producing no perturbing effect.

ROTHE, *Spontaneous Current and Velocity Distribution for Oxide Cathodes* (4.210), from measurements on technical triode valves, employing oxide-cathodes, obtained further evidence for obedience to Maxwell's Law, but with the average velocity 1·5 to 2·2 times that which would be expected on the kinetic theory, from the known cathode temperature.

MOLLER and DETELS, *Determination of Filament Temperature in Electron Tubes* (4.220), investigated the possibility of using the distribution of velocities of the emitted electrons to determine the temperature of the emitting filament, by plotting the logarithm of the emission current against the retarding potential V. They determined by calculation the approximate point at which the linear relation between log i and V should break down. Measurements on tungsten and oxide cathodes gave results in agreement with their theory.

The fact that the initial velocities of the emitted electrons are not zero implies that the 3/2 law relating current and anode potential at constant temperature, in a vacuum tube operating under space charge conditions, will break down for low anode potentials. The deviations from the law were used by KATO, *The Effect of Initial Velocity of Electrons upon the Anode Current of a Vacuum Tube* (4.230), to determine the effective average initial velocity of the electrons. For filament temperatures ranging from 2,250° K. to 3,000° K. the effective velocity was found to change from 0·5 to 4 volts. The effective velocity measured by this method was shown to be independent of the dimensions of the electrodes.

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A comprehensive paper dealing with the problem of electron velocity distribution was published by DEMSKI, *Experimental Proof of Maxwell's Velocity Distribution Law for Electrons Emerging from a Hot Cathode* (4.300). DEMSKI is concerned particularly with tracking down errors in earlier measurements which gave contradictory results, the velocity distribution temperature being usually too high. DEMSKI discusses the errors due (*a*) to potential fall along the emitting filament (*b*) to temperature variations along the filament and (*c*) to the electric and magnetic fields of the heating current in measurements without an interrupter. Owing to the difficulties in using mechanical commutators, A.C. interrupters were preferred. DEMSKI established Maxwell's Law for the electrons emitted by tungsten and by oxide-coated filaments after and also during the forming process. In applying Maxwell's Law as a method of temperature measurement, DEMSKI recommends the use of a cathode in which the heating current is so arranged that the emitting surface is an equipotential, and the net magnetic field external to the cathode equal to zero.

Some measurements of the electron velocity distribution from thoriated tungsten filaments were made by REYNOLDS, *Schottky Effect and Contact Potentials of Thoriated Tungsten Filaments* (4.310). He found Maxwell's Law verified for high electron velocities, but observed anomalous effects with small retarding potentials.

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SECTION 5

EFFECT OF APPLIED ELECTRIC FIELD AT THE SURFACE OF THE EMITTER (SCHOTTKY EFFECT)

ACCORDING to Richardson's original simple theory of thermionic emission, the current between the emitting cathode and the anode should be independent of the voltage between anode and cathode provided this voltage is sufficiently high to prevent any electrons escaping to the tube walls. As SCHOTTKY, *Influence of Structure Effects in particular, the Thomson Image Force on the Electron Emission of Metals* (5.100), first showed, however, true saturation is never attainable because the number of electrons escaping from the emitter varies with the intensity of the applied electric field at the surface. Schottky's explanation (see 1.140) based on the image force theory of the work function led to the following expression for the thermionic emission current i when the applied surface field equals $\frac{dV}{dx}$ and i_0 is the emission current for $\frac{dV}{dx} = 0$.

$$i = i_0 \exp \left\{ e^{3/2} \left(\frac{dV}{dx} \right)^{\frac{1}{2}} / kT \right\}.$$

The effect of increasing $\frac{dV}{dx}$ is thus to increase slightly the emission current i. Preliminary experiments by SCHOTTKY working with the emission from tungsten gave results in agreement with the theory in the range of surface fields considered.

The theoretical derivation of the formula for the effect was considered from a more general point of view by SCHOTTKY (see 1.243). Incidental confirmation of the result was obtained by DUSHMAN and his collaborators in their work on the thermionic emission constants of pure tungsten (see 2.260).

An extended investigation of the validity of Schottky's formula was carried out in 1928 by DE BRUYNE, *The Action of Strong Electric Fields on the Current from a Thermionic Cathode* (5.200). Working with tungsten filaments of diameters ranging from 0·002 to 0·01 cm., and at temperatures from about 1,600° to 2,000° K., the Schottky result was tested and found correct for surface fields at the emitter from 20,000 to 10^6 volts/cm. It was found, however, that for the largest diameter filament (0·0104 cm.) the Schottky

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law apparently broke down for field strengths at the surface exceeding about 250,000 volts/cm. The discrepancy is explained as being due to irregularities on the surface of the filament, which give an effectively much greater surface field than that calculated for a smooth surface. Assuming the validity of the Schottky formula, the results enable a value for the electronic charge e to be deduced. This came out to be $e = 4{\cdot}84 \times 10^{-10}$ e.s.u.

In an investigation similar to that of DE BRUYNE, BARTLETT, *The Increase of Thermionic Currents from Tungsten in Strong Electric Fields* (5.210), found that, although at a given temperature the logarithm of the emission current varied approximately in proportion to the square root of the surface field (as predicted by the Schottky relation), there was evidence of a slight systematic discrepancy. Furthermore, the slope of the log i against $\left(\frac{dV}{dx}\right)^{\frac{1}{2}}$ graph obtained for different temperatures, did not vary inversely as the temperature as required by SCHOTTKY, and equalled the value to be expected by Schottky's formula at only one temperature (1,900° K.). BARTLETT used filaments of diameter 0·08 and 0·095 mm. respectively, and his measurements extended over the temperature range 1,500° to 2,400° K. and up to surface fields of 60,000 volts/cm. The reason suggested by BARTLETT for the observed departures from Schottky's law is that in deriving the law the image force on the escaping electron is alone considered, whereas actually there must be a negative space charge effect also coming into account, due to the large number of electrons which emerge from the surface only to be turned back by the image force.

PFORTE, *Increase of the Saturation Current from a Thermionic Emitter by Intense Electric Fields* (5.220), measured the change in saturation current from a tungsten filament for field strengths at the surface up to $0{\cdot}7 \times 10^6$ volts/cm. and for temperatures ranging from 1,400° to 2,100° K. His results agreed with de Bruyne's in confirming Schottky's formula both as regards linear variation of log i with $\left(\frac{dV}{dx}\right)^{\frac{1}{2}}$ and agreement between the slope of the log i against $\left(\frac{dV}{dx}\right)^{\frac{1}{2}}$ line as measured and as computed for the different temperatures worked with.

VAN VELZER and HAM, *Thermionic Emission from Tungsten and the Schottky Equation* (5.230), state in a brief note that using extended high-voltage clean-up of the tungsten filament, their measurements confirm Schottky's relation for fields up to 10^6 volts/cm., there being apparently no sign of an upper limit beyond which the formula would not apply.

Assuming the validity of Schottky's formula, HAM, *The Use of the Schottky Relation in the Determination of the Thermionic Work Function of Tungsten* (5.300), drew attention to a method of

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measuring the work function Φ. Taking as the emission formula (zero applied field) $i_0 = A \exp(-\Phi e/kT)$ and introducing Schottky's relation $i = i_0 \exp\left(e^{3/2}\left(\frac{dV}{dx}\right)^{\frac{1}{2}}/kT\right)$ it follows that

$$\log i = \log A - \left[\Phi e + e^{3/2}\left(\frac{dV}{dx}\right)^{\frac{1}{2}}\right]\frac{1}{kT}.$$

By determining i for a fixed temperature and for different values of $\left(\frac{dV}{dx}\right)$ and then plotting i against $\left(\frac{dV}{dx}\right)^{\frac{1}{2}}$ a straight line of slope S and intercept I on the log i axis is obtained, where $\log A - (\Phi e/kT) = I$ and $(e^{3/2}/kT) = S$. Clearly $I = \log A - \Phi S/e^{\frac{1}{2}}$. Repeating the procedure for different temperatures and plotting corresponding values of I and S, a straight line of slope $(-\Phi/e^{\frac{1}{2}})$ will be obtained if Φ is independent of temperature, i.e. if the emission equation $i_0 = A \exp(-\Phi e/kT)$ is correct. If deviations from a straight line occur these can be used to determine the variation of Φ with temperature and the correct form of the emission equation can be

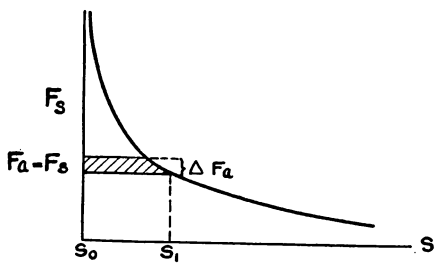

FIG. 5.400*a*.

determined. An advantage of the method is that precise temperature measurements are not required and the values of $\left(\frac{dV}{dx}\right)$ need only be relatively accurate.

The problem of the effect of the applied field on the emission from composite surfaces was considered by BECKER and MUELLER, *Electrical Fields near Metallic Surfaces* (5.400), who pointed out that with pure tungsten or tungsten completely covered, for example, with cæsium-on-oxygen, saturation is readily obtainable, but with composite surfaces, for example, tungsten partially covered with cæsium-on-oxygen, saturation is not obtained. This lack of saturation can be shown not to be due to space charge effect or ionisation of gas. It must be due to the adsorbed electro-positive atoms partially covering electro-negative surfaces. Schottky's explanation of the slight lack of saturation in the case of clean surfaces is entirely inadequate when applied to composite surfaces. Representing by F_s the surface field, that is, the total force acting on an escaping electron from all causes, including the image force, but excluding the externally-applied electric field F_a, the curve in Fig. 5.400*a*

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shows in a general way how F_s varies with distance away from the surface. It is clear from the diagram that if the applied field is increased to $F_a + \Delta F_a$ the work function will change by $\Delta\Phi = -\Delta F_a s$, and hence $\frac{d\Phi}{dF_a} = -s$. Putting $i = AT^n \exp(-\Phi e/kT)$

$$\log_{10} i = \log_{10} A + n \log_{10} T - \Phi e/2{\cdot}3kT$$

and

$$\frac{d \log_{10} i}{dF_a} = -\frac{e}{2{\cdot}3kT}\frac{d\Phi}{dF_a} = \frac{e}{2{\cdot}3k}\frac{s}{T} = \frac{11{,}600}{2{\cdot}3}\cdot\frac{s}{T}.$$

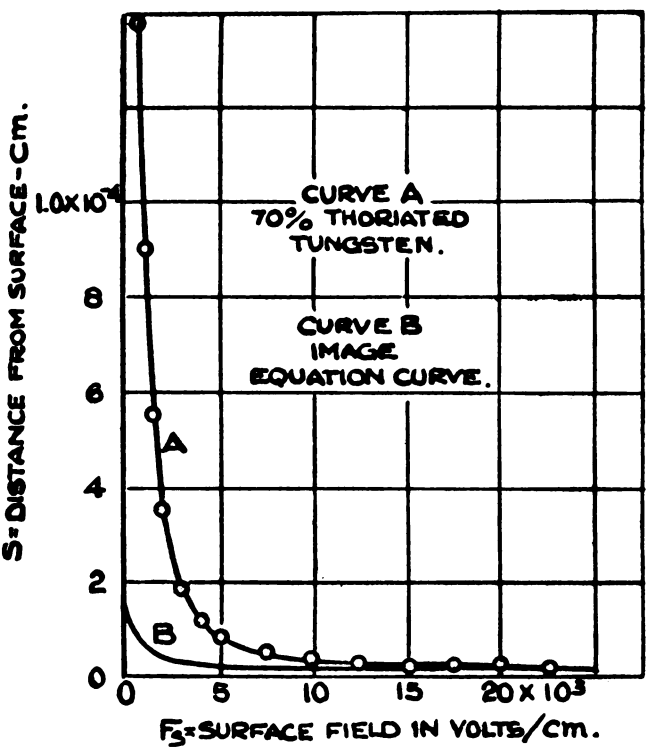

FIG. 5.400*b*.

Thus by determining $\frac{d \log_{10} i}{dF_a}$ experimentally, the value of the distance s is determinable at which $F_a = F_s$, i.e. F_s, is determinable as a function of s.

BECKER and MUELLER apply this method to find the surface field for a 70 per cent. thoriated tungsten surface (see Curve A, Fig. 5.400*b*). The surface field is regarded as the sum of a Schottky image force field (Curve B, Fig. 5.400*b*) and a field due to the layer of thorium atoms. It appears by subtracting curve B from curve A

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that the adsorbed thorium ions produce a field which, close to the surface, is very large and in a direction to help electrons escape. At larger distances the field is in the opposite direction and may be appreciable as far out as 2,500 atom diameters. The huge fields near the surface are responsible for the decreased work function, whilst the reverse fields farther out account for the marked lack of saturation at ordinary applied potentials. The present work throws doubt on the correction for Schottky Effect which DUSHMAN and EWALD applied in their determinations of the Richardson constants of thoriated tungsten. BECKER and MUELLER consider it not impossible that, when suitable account is taken of the effects dealt with in their paper, thoriated tungsten may give the universal value 60·2 for the constant A.

Becker and Mueller's method represented a new mode of approach to the Schottky effect. Instead of assuming the surface field to be that given by the image force theory, calculating the effect of applied field and comparing with experiment, the experimental data are now used to find the actual form of the surface field, which is then compared with that given by the image force theory.

LAURITSEN and MACKEOWN, *Electric Fields near the Surface of Tungsten Wire of Small Diameter* (5.401), worked with a tungsten filament of 0·00156 cm. diameter and measured the thermionic current for applied field strengths up to 2×10^6 volts/cm. They found that for distances greater than 3×10^{-7} cm. the surface field is given accurately by the image force equation, but is less for smaller distances. Increase of temperature also caused a decrease in the field strength very near the surface.

Becker and Mueller's measurements on thoriated tungsten were repeated and extended by REYNOLDS, *Schottky Effect and Contact Potentials of Thoriated Tungsten Filaments* (5.410), who determined the relation between thermionic emission and voltage over a range of anode potentials from 10 volts retarding to 500 volts accelerating, the latter corresponding to an applied field at the surface of the emitter of nearly 500,000 volts/cm. For sufficiently high applied fields (greater than 10,000 volts/cm.) Schottky's formula was verified, but the formula ceased to be valid for lower field strengths, even for fully activated surfaces. The critical field strength (10,000 volts/cm.) appeared to be independent of the proportion of thorium on the surface ($\theta = 0{\cdot}3$ to $\theta = 1{\cdot}0$) and of the temperature ($T = 1{,}100°$ K. to $1{,}600°$ K.). The apparent lack of saturation at low fields was accentuated by positive ion bombardment, which probably resulted in surface roughening and increased local fields. A patch theory of the thoriated tungsten surface is put forward to explain the lack of saturation at applied fields below a limiting value.

The form of the surface field near a thoriated tungsten surface was investigated by a variant of Becker and Mueller's method, using the shift of the photo-electric long wave limit, by LINFORD,

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Electrostatic Surface Fields near Thoriated Tungsten Filaments by a Photo-electric Method (5.420). The shift in the long wave limit with increasing applied electric field was determined. The surface field derived from the measurements approximated to the image force field at distances less than $1 \cdot 5 \times 10^{-6}$ cm. The surface field agreed very nearly with that obtained by BECKER and MUELLER for 70 per cent. thoriated tungsten from thermionic measurements. LINFORD considered this agreement as showing definitely that sufficiently large fields are present near thoriated tungsten surfaces to cause the known deviations from Schottky's formula. A "patch" theory is suggested to explain the surface fields.

An outline of a possible "patch" theory explanation of the saturation effects observed with composite surfaces was given by BECKER in a brief note, *The Ion Grid Theory of the Decrease in Work Function for Composite Surfaces* (5.430). The theory assumes that when electro-positive ions are adsorbed on electro-negative surfaces some of the atoms are ionised, forming a positively-charged grid

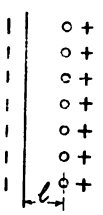

FIG. 5.430*a*.

very close to the electrode (Fig. 5.430*a*). For a large plane surface the positive grid together with its negative image in the surface produces strong fields close to the surface, but only weak fields at large distances. In this case the work function is reduced by $4\pi\sigma l$ and the emission current saturates just as well as for clean surfaces. In most experiments, however, the surface consists of irregularly-oriented facets or else the ions form clusters, owing to the presence of electro-negative gases. The distribution of the fields then depends greatly on the size of the facets or clusters. Computation shows that for small applied fields the work function is decreased by a fraction f of $4\pi\sigma l$, but as the field increases this fraction increases rapidly until for sufficiently high field strengths, $f = 1$ and saturation equally good as for clean surfaces is obtained.

A critical discussion of various "patch" theories put forward to explain departures from Schottky's equation is given by COMPTON and LANGMUIR, *Electrical Discharge in Gases* (5.440). These writers consider a chessboard pattern of alternating free and thorium-coated squares on a tungsten surface, and evaluate the variation of emission with field strength. They find that, although the theory predicts variations of emission with field strength, which are qualitatively

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in the right direction, closer examination shows that the theoretical variations are quite unlike those actually observed. In the first place, to obtain departures from Schottky's formula comparable with the observed effects, the patches must be assumed to contain many thousands of atoms, which is a very improbable assumption. In the second place, the patch theory predicts a departure from Schottky's law, which is small with small fields and increases with large fields, whereas actually the reverse is the case. These difficulties are also shown to be inevitable with any theory based on the assumption of local inhomogeneities at the surface.

COMPTON and LANGMUIR proceed to examine the form of theory suggested by BECKER, which they term the adion field theory (see 5.430). It is found that this theory gives qualitatively the observed departures from Schottky's formula, but, like the patch theory, it fails to account for the fact that the departure is greatest for weak fields. Both the patch theory and the adion field theory are explanations based on changes in the work function, A being assumed constant. NORDHEIM and FOWLER (1.500 and 1.520) showed that A is not in general a constant, particularly for activated surfaces, and it may be, therefore, that for activated surfaces A depends on the field in such a way as to account for the abnormally small emission in weak fields.

The attempt to explain the thermionic work function as due to the space charge field of electrons streaming out from the surface and then being turned back by the field was taken up by BARTLETT, *A Space Charge Interpretation of Thermionic Work Function* (5.450). To account on this theory for the increase in thermionic current with strong fields it is necessary for the work function field to extend so far out that an escaping electron is much nearer to its neighbours than to the surface. Assuming a Maxwellian distribution of electron velocities right down to the surface, a suitable value for the work function can be obtained, the increase of current with strong fields deduced on the theory is stated to be in better agreement with experiment than Schottky's formula based on the image force. A first order correction to allow for the Fermi-Dirac statistics, which must be effective as the interior of the metal is approached, improves the agreement, although involving too large a variation of work function with temperature. BARTLETT considers that a strict application of Fermi-Dirac statistics should remove this difficulty and should also provide an explanation of the cold electron emission.

WATERMAN, *Density Distribution of Electron Gas in Equilibrium with a Hot Body* (5.451), working on the same lines as BARTLETT, considered the variations of potential and electron concentration with distance from a plane emitting surface using Poisson's equation, connecting the electro-static potential and the electric charge density, and the Fermi-Dirac statistics. Among other conclusions, WATERMAN states that the image force explanation of the work

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function is not applicable, and that the space charge effect in the surface is quite adequate to give the magnitude of the work function.

A possible explanation of the departures of the constant A from the theoretical universal value, and of the anomalously large values of the Schottky effect, which are most marked for surface films, is outlined by KINGDON, *Thermal Fluctuations of the Surface Potential of a Cathode as affecting Thermionic Emission* (5.460). KINGDON considers covered patches of the surface to represent minute resistance-with-capacity systems across which a fluctuating potential is generated by the thermal agitation of the electrons. The radius of each patch is assumed to equal the critical distance from the surface which an electron must reach in order to escape. This varies with the applied surface field and hence the patch radius and the derived fluctuating potential varies. The corresponding changes in the thermionic emission are found to be related to the applied field in a way similar to that actually observed. The introduction of the surface resistance between the patch and the underlying surface may be made to account for variations in the constant A by assuming a suitable temperature coefficient for the surface resistance.

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SECTION 6

THE PHOTO-THERMIONIC EFFECT

IT was pointed out by CASE, *New Strontium and Barium Photo-electric Cells* (6.100, 6.101), that the saturation current in an audion tube containing a filament coated with alkaline earth oxides, is increased by illumination of the filament. CASE attributes the effect to photo-electric emission of alkali earth metals formed on the filament by reduction of the oxides. The effect was further investigated by MERRITT, *Photo-electric Phenomena in Coated Filament Audion Bulbs* (6.110), who found that the increased current obtained increased with the filament temperature, although not so rapidly as the "dark" thermionic current. Ascribing the increased current to photo-electric effect, a remarkable increase of photo-electric activity with temperature is indicated. This conclusion is questioned by ARNOLD and IVES, *The Growth and Decay of Photo-thermionic Currents from Oxide-coated Filaments* (6.120), who from a study of the growth and decay curves of the increased current suggest that the effect of light is to change the character of the emitting material in some way (a similar process to the effect of light on the resistance of selenium).

The existence of an appreciable time lag before the increase of current, following upon the illumination of the cathode in an audion, was questioned by MAJORANA, *Two New Experiments with the Audion* (6.121). He used flickered light at a frequency of 1,500 flashes a second and detected the increased current in a telephone current. He concluded that an appreciable effect is produced immediately the light falls on the cathode. In a second experiment MAJORANA focussed flickered light on different parts of a bismuth emitter, obtaining different responses depending on the orientation of the emitter. Other metals failed to show similar effects.

Investigations of thermionic tubes, in which photo-electrically sensitive metals had been introduced, are described by TYKOCINER and KUNZ, *A New Photo-electric Cell* (6.130), *Photo-electric Cells with Hot Filaments* (6.131), HYATT, *The Modification of the Thermionic Current in Vacuum Tubes when Potassium Deposited on the Inside Walls or Grid of the Tube is Illuminated* (6.132), and ALBERS, *A New Photo-electric Valve* (6.133), *An Investigation of a Photo-electric Valve coated with Potassium* (6.134). By illuminating such tubes the current-voltage characteristics were shown to be modified in various ways.

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The effect on the anode current of concentrating the light from an arc on to a tungsten filament was examined by HENRIOT and MOENS, *Action of Light on Thermionic Effect* (6.140). An increased emission was observed which the authors show to be due to an elevation in the temperature of the filament caused by the incident light and not to a photo-electric effect properly speaking.

In NEWBURY'S work, *The Effect of Light on the Electron Emission from Oxide-coated Filaments* (6.150), on the photo-thermionic effect, light from a quartz arc was focussed on an oxide-coated filament and the increased thermionic current measured. For one filament the change of the saturated thermionic emission was as shown in the accompanying Table :—

Filament current.	Thermionic emission.	Increase due to light.
Amps.	Micro-amps.	Micro-amps.
2·5	3·5	0·16
2·7	7·0	0·26
2·9	13·8	0·40
(appr. 1,000° C.) 3·1	28·5	0·48
3·3	46·0	0·38
3·5	73·0	0·35
(appr. 1,200° C.) 3·7	173·0	0·15

Measurements were also made of the variation of the additional current with the wave-length of the light for the hot filament (1,000° and 1,200° C.) and the cold filament (21° C.). The same shape curve was obtained, whether the filament was hot or cold, but the absolute magnitude of the effect for the cold filament was only 1/300th that for the hot filament. It is claimed that the effect cannot be attributed to the addition of an abnormally high photo-electric emission to the thermionic emission. The explanation indicated assumes a normal photo-electric emission which appreciably modifies the factors controlling thermionic emission.

The photo-thermionic effect for filaments coated with cerium oxide, thorium oxide, calcium oxide or calcium iodide was examined by BERGER, *The Effect of Light on the Electron Emission from Cerium Oxide* (6.160), who found in each case a positive effect, but in no case so large as that given by a " Western Electric " filament. In a special study of the emission from cerium oxide, BERGER found the amount of the extra current independent of the metal forming the core of the coated filament, thus showing the effect to be a property of the coating. At 980° C. the photo-thermionic current was about double the true thermionic emission, but at 1,080° C. the true thermionic effect greatly exceeded the photo-thermionic. Addition of small quantities of air, hydrogen or oxygen decreased

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the effect of light on the emission. BERGER considered his experiments inadequate to decide whether the observed increases represented a true photo-electric effect or whether they arose from a change in surface conditions.

BODEMANN, *Change in the Thermionic Emission of Oxide-coated Metal Filaments due to Illumination with Ultra-violet Light* (6.170), found for oxide-coated platinum foil, when illuminated with the ultra-violet light from a mercury vapour lamp, an additional thermionic current, which with suitable anode voltage was some 200 times the photo-electric current given by the foil when cold. The additional current depended on the anode voltage employed and tended to zero when this was sufficiently high to produce complete saturation. Repeated and prolonged out-gassing also produced a steady decrease in the magnitude of the additional current, reintroduction of gas restoring the higher values. BODEMANN concludes from his results that when gas is present in the surface layer the effect of illumination is to liberate electrons loosely attached to the gas molecules or atoms, thereby reducing negative space charge. The additional current, according to this view, is a part of the true thermionic current, which in the unilluminated condition is held back by negative space in the surface, the effect of illumination being to disperse the space charge.

The question whether the photo-thermionic effect observed with tungsten filaments is a direct light effect or merely an increase in the thermionic emission due to rise in temperature was taken up again by DEAGLIO, *Action of Light on Thermionic Phenomena* (6.180). His measurements showed that at low anode voltages (i.e. too low to produce saturation), the effect is much larger—some hundred times—than the maximum possible effect attributable to thermal action. With sufficiently high anode voltage to produce saturation, DEAGLIO found, as observed by BODEMANN for oxide-coated filaments, a much smaller photo-thermionic effect, which was of the same order of magnitude as the expected thermal effect, and under these conditions no true light effect could be established.

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SECTION 7

THORIATED FILAMENTS AND OTHER THIN FILM EMITTERS

PRELIMINARY announcements of the discovery of the enhanced emission obtainable from tungsten filaments containing a trace of thoria were made by LANGMUIR as early as 1914, and are referred to in Section O (0.220, 0.223). The first detailed description of the phenomena, however, was not given by LANGMUIR until 1923. LANGMUIR, *The Electron Emission from Thoriated Tungsten Filaments* (7.100, 7.101), found that, using a filament containing a little thoria (ThO_2, 1 to 2 per cent.), on flashing for three minutes at 2,800° K. and then lowering the temperature to between 2,000 and 2,100° K. for a suitable activating period, the filament acquires an abnormally high emission at the testing temperature (1,400° to 1,500° K.) many times greater than that of a tungsten filament at the same temperature, which persists so long as the temperature is kept below 1,900° K. Heating the filament at a temperature in the range 2,200° to 2,600° K. resulted in deactivation, i.e. the loss of the abnormally high electron emission. Starting with a deactivated filament and taking measurements of the emission at the testing temperature after successive periods of activation at the activation temperature, LANGMUIR determined curves showing the rate of activation of the filament. Deactivation curves were determined in a similar fashion. LANGMUIR gives reasons, based on the activation and deactivation curves and on other evidence, for believing that the activation is due to the deposition on the tungsten of a layer of thorium one atom thick. He also suggests (without proof) that the fraction of the surface covered with thorium atoms will be given by

$$\theta = \frac{\log i - \log i_0}{\log i_1 - \log i_0}$$

where i_0 is the saturation current corresponding to pure tungsten ($\theta = 0$), i_1 is that from the fully activated surface, whilst i is the current when the fraction of the surface covered by thorium has the value θ. If b_0, b_1 and b are the values of the work function constant in the thermionic emission equation, corresponding to these three cases, and if the constant A is assumed to be unchanged by activation,

$$\theta = (b - b_0)/(b_1 - b_0).$$

At the deactivating temperature all thorium evaporates from the surface as rapidly as it arrives there by diffusion. At the activation temperature the rate of activation $\frac{d\theta}{dt}$ depends on the difference

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between the rate of diffusion to the surface and the rate of evaporation from the surface. LANGMUIR proves that back diffusion of thorium into the tungsten does not occur. In order to explain the observed shape of the activation curve, LANGMUIR differentiates between two kinds of evaporation of thorium atoms :—

(1) *Induced evaporation of thorium atoms.*—This will occur when a thorium atom diffuses up to the surface at a place where there is already a thorium atom. If, as established by Langmuir, the original thorium atom is in every case expelled, the rate of induced evaporation will equal $DG\theta$, where D is the coefficient of diffusion of thorium atoms through tungsten and G is the concentration gradient of thorium atoms at the surface.

(2) *Normal evaporation of thorium atoms from the tungsten substratum.*—If the amount of normal evaporation is small compared with the rate of diffusion of atoms to the surface, the limiting condition of the filament will be with $\theta = 1$. This is the case for not too low a concentration gradient G. At higher temperatures, however, normal evaporation comes into play, and the fraction of the surface covered will vary with temperature. At sufficiently low temperatures the thoriated filament will function like a pure metal filament, obeying Richardson's equation, but with a lower value of the constant b. More generally, the emission will be governed by the equation

$i = 60{\cdot}2\ T^2 \exp(-b/T)$ amps. cm.$^{-2}$ with $b = b_0 + (b_1 - b_0)\theta$, where $b_0 = 52{,}600$ and $b_1 = 34{,}100$, are the values for pure tungsten and fully activated ($\theta = 1$) thoriated tungsten. In the following table, taken from Langmuir's paper, the equilibrium value of θ and the corresponding emission current are worked out for temperatures ranging from 1,300° K. to 3,000° K.

Properties of a thoriated tungsten filament in the steady state.

(Filament contains 1 per cent. ThO_2.) (Diameter = 0·00389 cm.)

T° K.	θ.	*i* (amps./cm.²).	Life (hours).
1,300	0·99997	$4{\cdot}14 \cdot 10^{-4}$	—
1,400	·99975	$3{\cdot}12 \cdot 10^{-3}$	—
1,500	·99878	0·0179	—
1,600	·99528	0·0812	—
1,700	·9848	0·287	—
1,800	·9605	0·772	720,000
1,900	·9191	1·59	94,000
2,000	·8713	2·89	15,100
2,100	·781	3·43	2,897
2,200	·551	1·24	643
2,300	·139	0·114	164
2,400	·0601	0·168	47
2,500	·355	0·357	14·6
2,600	·0207	0·774	5·01
2,800	·0088	3·48	0·74
3,000	·0041	13·5	0·14

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The "life" given in the fourth column is the time required for the thoria content to be reduced to $1/e$ of its original value. The steady emission from a thoriated filament is seen from the table to reach a maximum somewhat above 2,000° K. and then fall gradually to that of pure tungsten.

Other data determined by LANGMUIR from the analysis of his experimental results are the following :—

$$\log_{10} D = 0{\cdot}044 - 20{,}540/T \quad (D \text{ in cm.}^2/\text{sec.}).$$

Rate of diffusion of Th atoms to the surface = DG (G in atoms/cm.^{4}).

Rate of normal evaporation E_n of thorium atoms from a partly covered tungsten surface ($\theta = 0{\cdot}2$ to $0{\cdot}8$) (E_n in atoms/cm.^{2}/sec.)

$$\log E_n = 31{\cdot}43 - 44{,}500/T.$$

Rate of induced evaporation E_i :—

$$E_i = DG\,(0{\cdot}82\,\theta + 0{\cdot}18\,\theta^3).$$

Rate of activation

$$N_0 \frac{d\theta}{dt} = DG - E_n - E_i = DG\,(1 - 0{\cdot}82\,\theta - 0{\cdot}18\,\theta^3) - E_n,$$

where N_0 is the number of thorium atoms per square cm. of surface for a saturated film ($\theta = 1$). N probably equals $0{\cdot}756 \times 10^{15}$.

Heat of reduction of thoria in tungsten = — 138,000 cal. (absorbed).

Heat of evaporation of Th from tungsten = 204,000 cal.

Heat of diffusion of Th through tungsten = 94,000 cal.

The effect of bombardment of the thoriated tungsten surface, with positive ions of different gases, was investigated by KINGDON and LANGMUIR, *The Removal of Thorium from the Surface of a Thoriated Tungsten Filament by Positive Ion Bombardment* (7.110). They found that H ions produced no sputtering even up to 600 volts energy. *Ar*, *Cs*, *Hg* and *He* ions all started to sputter the thorium atoms at about 50 volts energy. The number of impacting ions of 150 volts energy per sputtered thorium atom varied from 12 for *Ar* and *Cs* to 45 for *Ne* and 7,000 for *He*. The rate of sputtering was found to be greater when 0·95 of the surface was covered with thorium than when it was completely covered. It is suggested that, for the removal of the first few atoms, two successive impacts on the same thorium atom are necessary.

KINGDON and LANGMUIR, *Thermionic Effects Caused by Alkali Vapours in Vacuum Tubes* (7.120, 7.121), investigated the thermionic emission from heated filaments in cæsium vapour. They found that with metallic cæsium in the tube at 25° C. the cæsium vapour forms an adsorbed film, consisting of a single layer of atoms, covering the surface of the heated tungsten filament, even at filament temperatures of 600° K. or more.

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The stability of the film is attributed to the fact that the electr affinity or work function of the tungsten surface (4·52 volts) is l than the ionisation potential of the cæsium atom (3·90 volts), that a cæsium atom striking the surface is robbed of its valer electron and, in the form of a positive ion, is held to the tungs surface by the image force. The presence of the cæsium film made evident by the very high electron emission produced and measurement of the contact potential with respect to a cæsium f tungsten filament.

Adsorbed films of electro-negative gases, one atom thick, w also formed on tungsten and decreased the electron emission, but admitting cæsium vapour the cæsium atoms formed a second la on top of the gas atoms and were retained by the latter at hig temperatures than when in direct contact with the tungst Cæsium films stabilised in this way remained intact up to tempe tures of about 900° K. and emitted at this temperature saturat currents of the order of 0·3 amps./cm.². At temperatures bel those at which evaporation of the cæsium was appreciable emission current conformed to the equation $i = 60 \cdot 2\ T^2\ e$ $(-b_o/T)$, with $b_o = 16{,}000$, corresponding to $\phi = bk/e = 1$ volts. The same value was obtained whether the cæsium v adsorbed directly on the tungsten or on an adsorbed electro-negat film.

For a surface partially covered with cæsium the work funct (or electron affinity) of the surface, ϕ is assumed to be linea related to the work function of the pure tungsten and fully cæsia surfaces, ϕ_w and ϕ_{cs} respectively, i.e. $\phi = \phi_w - \theta\ (\phi_w - \phi$ where θ is the fraction of the surface covered. For $\theta = 0 \cdot$ ϕ equals 3·90 volts, which is the ionisation potential of cæsiu Cæsium leaving the surface with $\theta > 0 \cdot 20$ (α state of the surfa would be expected to be in the form of neutral atoms and $\theta < 0 \cdot 20$ (β state of the surface) in the form of positive ions. T change-over from the α to the β surface state was observed exp mentally at a temperature in the neighbourhood of 1,150° K. the β state every atom striking the surface leaves as a positive and the saturation positive current obtainable from the filam is related to the pressure of the cæsium vapour by the equat

$$i = \frac{ep}{\sqrt{2\pi MkT}} = 0 \cdot 367\ \frac{p}{\sqrt{T}}\ \text{amps/cm.}^2,$$

where T is the tempe ture of the bulb, M the mass of a cæsium atom, k Boltzman constant, and e the electronic charge. The positive ion current thus independent of filament temperature provided this is above t critical temperature, and is the same for any filament mater giving the effect, and is proportional to the vapour pressure. Th conclusions were confirmed experimentally.

Using a fully thoriated filament in place of the pure tungst filament, the enhanced electron emission and the generation

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positive ions on introducing cæsium vapour were not obtained, in agreement with the fact that the work function of the thoriated surface 2·94 volts is less than the ionisation potential of the cæsium, 3·90 volts.

Investigations of the electron emission from adsorbed films of oxygen, thorium and cæsium on tungsten are described by KINGDON, *Electron Emission from Adsorbed Films on Tungsten* (7.130). To maintain a complete film of oxygen on tungsten, oxygen gas had to be supplied continuously, the tungsten filament being gradually consumed by oxidation. The emission corresponded to $A = 5 \times 10^{11}$ amps./cm.^{2} deg.^{2}, $b = 107{,}000$ $\{i = AT^2 \exp(-b/T)\}$. A study of the emission from thoriated filaments of various degrees of activation, gave a series of pairs of values of A and b. Assuming a linear relation between θ and b, as conjectured by LANGMUIR, $b_\theta = \theta b_{th} + (1 - \theta) b_w$, the value of θ could be computed and the relation between A and θ determined. It was found that A decreased from $A = 60$ for $\theta = 0$ to a minimum at $\theta = 0{\cdot}8$, rising slightly to $A = 7{\cdot}0$ for $\theta = 1$. The empirical formula $A = [7^\theta + 60^{1-\theta} - 1]$ fitted the results.

This formula is generalised, writing a_1 and a_2 for the numerical values of A for the complete adsorbed film and for the underlying layer respectively,

$$A_\theta = [a_1{}^\theta + a_2{}^{(1-\theta)} - 1]\, A_o,$$

where A_o is the dimensional unit 1 amp./cm.^{2} deg.^{2}. The complete emission equation then runs

$$i = [a_1{}^\theta + a_2{}^{(1-\theta)} - 1]\, A_o T^2 \exp\{-[b_1 o + b_2 (1 - \theta)]/T\}.$$

This type of formula was also found to account satisfactorily for a number of effects observed with cæsium, and cæsium on oxygen films. For the latter complete films could not be obtained, but the results indicated that A_{CsO} was less than 0·003 and b_{CsO} a little less than 8,300. Substituting these values and the values for the oxygen film in the equation given, the emission should have a maximum at $\theta = 0{\cdot}993$ instead of at $\theta = 1$. Various experiments on the effect of varying the cæsium vapour pressure and the effect of evaporating Cs from the surface of the filament confirmed this conclusion. For cæsium films on tungsten the emission was computed to be a maximum for $\theta = 0{\cdot}987$. Addition of nitrogen to a fully thoriated filament increased the emission about five-fold at 1,400° K., an effect attributed to a change in the A value.

Thermodynamic considerations involving the Saha equation for thermal ionisation were applied by LANGMUIR and KINGDON, *Thermionic Effects caused by Vapours of Alkali Metals* (7.140), to show that there must be a relation between the positive ion emission from a heated filament in the vapour of an alkali metal and the electron emission. For high filament temperatures and low pressures of the vapour the electron emission is the same as in the

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absence of vapour (i.e. no alkali metal coating is formed), and knowing the electron emission from the filament and the Saha equilibrium constant for the vapour, the positive ion emission was computed and found to agree with that observed experimentally, using oxygen-coated, thorium-coated and uncoated tungsten filaments in cæsium vapour. At lower filament temperatures the

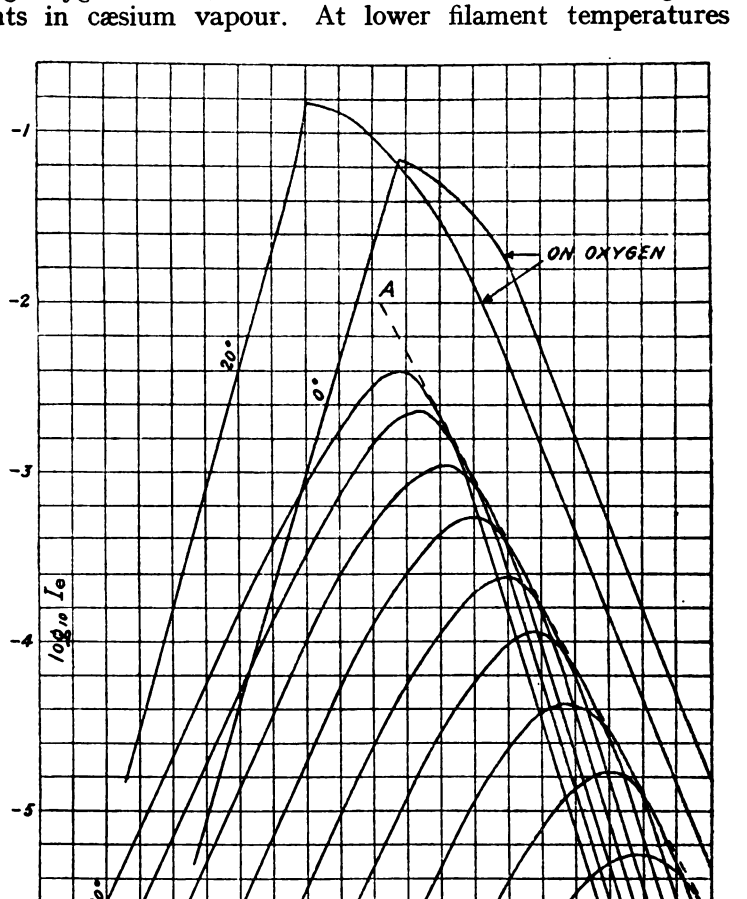

FIG. 7.140*a*.—*Electron emission from tungsten filaments in cæsium vapour at various pressures.*

electric image force causes a fraction θ of the filament surface to be covered by a layer of adsorbed ions which share electrons with the underlying metal.

Emission data based on some 220 determinations of the electron emission in cæsium vapour are given by LANGMUIR and KINGDON

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as a series of curves, Fig. (7.140*a*). The ordinates give $\log_{10} I_e$, where I_e is the emission current in amps./cm.^{2} and the abscissæ are the reciprocals of the absolute temperatures of the filament, multiplied for convenience by 1,000. The temperature of the bulb, which fixed the vapour pressure of the cæsium vapour, is given in degrees Centigrade on each curve. Two of the curves refer to filaments covered with a complete layer of oxygen atoms before the admission of cæsium vapour. The rest were obtained with pure tungsten filaments.

At sufficiently low temperatures the filament was completely covered with cæsium atoms ($\theta = 1$) and the emission followed Richardson's Law giving the straight lines descending to the right in the figure (Region I). An intermediate region is reached and the current passes through a maximum (Region II). The current then decreases, $\log_{10} I_e$, again varying linearly with (1,000/T), this part of the curve corresponding to the loss of cæsium from the surface by evaporation, and to decreasing values of θ (Region III). The two straight line portions of the curves are reproduced by the formula

$$i = \alpha p^{\gamma} \exp(-\beta/T) \text{ amps./cm.}^2 \quad (p \text{ in bars}),$$

where α, β, γ are constants given in the following table :—

—	α.	β.	γ.
Cæsium on tungsten—			
Region I	$6{\cdot}65 \times 10^{-14}$	+ 31,300	− 0·70
Region II	$8{\cdot}6 \times 10^{-12}$	− 19,100	+ 1·66
Cæsium on adsorbed oxygen—			
Region I	104×10^{-10}	+ 23,400	− 0·62
Region III	$2{\cdot}95 \times 10^{-11}$	− 32,200	+ 1·96

The vapour pressure of cæsium at different temperatures, determined by measurements of the positive ion emission at a high filament temperature (1,500° K.), is given by the formula

$$\log_{10} p = 10{\cdot}65 - 3{,}992/T \quad (p \text{ in bars}).$$

LANGMUIR and KINGDON also give a theoretical discussion of the conditions in the cæsium film, both in the dilute state ($\theta < 0{\cdot}2$) and the concentrated state ($0{\cdot}2 \leqslant \theta < 1$).

Similar effects to these observed by LANGMUIR and KINGDON for tungsten filaments in cæsium vapour were investigated by KILLIAN, *Thermionic Phenomena caused by Vapours of Rubidium and Potassium* (7.141), using rubidium and potassium vapours. Figures 7.141*a* and *b*, which are analogous to Figure 7.140*a*, show the

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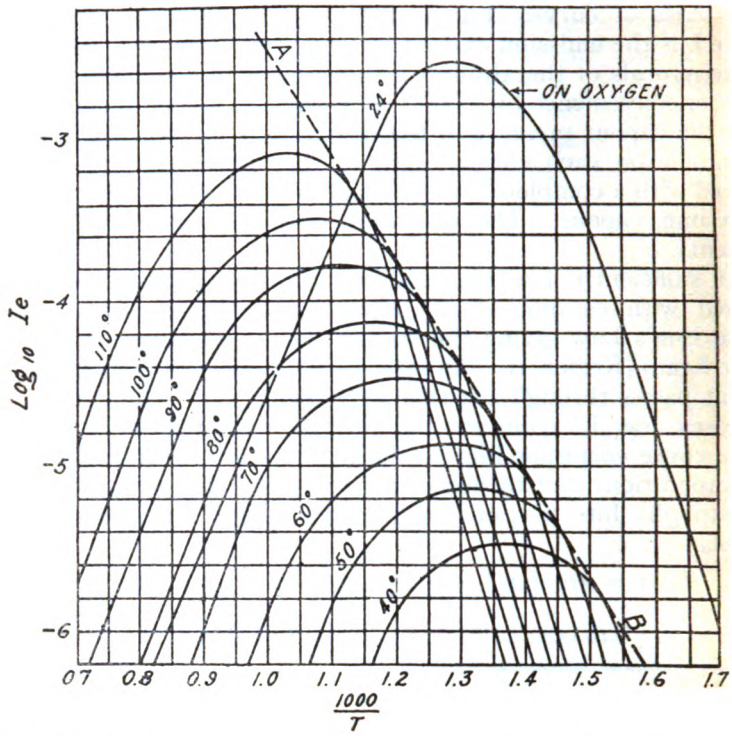

FIG. 7.141*a*.—*Electron emission in amperes per cm² from a tungsten filament in rubidium vapour at various pressures. Temperature of bulb in °C on each curve.*

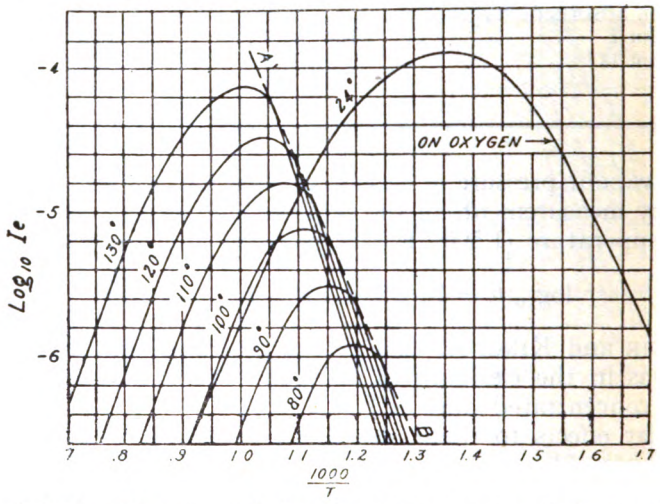

FIG. 7.141*b*.—*Electron emission in amperes per cm² from a tungsten filament in potassium vapour at various pressures. Temperature of bulb in °C on each curve.*

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results obtained in graphical form. For the constants α, β, γ in the equation $i = \alpha p^{\gamma} \exp(-\beta/T)$ KILLIAN found :—

Filament coating.	α.	β.	γ.
Rubidium on tungsten—			
Region I	$2{\cdot}86 \times 10^{-12}$	+ 32,000	− 1·12
Region III	$2{\cdot}08 \times 10^{-12}$	− 24,000	+ 2·52
Rubidium on adsorbed oxygen—			
Region I	—	+ 28,000	—
Region III	—	− 23,000	—
Potassium on tungsten—			
Region I	$5{\cdot}9 \times 10^{-9}$	+ 30,800	− 0·328
Region III	$1{\cdot}27 \times 10^{-13}$	− 25,600	+ 2·66
Potassium on adsorbed oxygen—			
Region I	—	+ 19,700	—
Region III	—	− 23,000	—

For the vapour pressures deduced from the positive ion emission at high filament temperatures, Killian's results lead to the formulæ

$$\text{Rubidium} \quad \log_{10} p = 10{\cdot}55 - 4132/T$$
$$\text{Potassium} \quad \log_{10} p = 11{\cdot}83 - 4964/T$$

KILLIAN also established from the experimentally determined value of the constant in the three halves power law for the space charge limited current, that in the positive ion emission each ion had a mass equal to that of a single atom.

The effect of cæsium vapour on the electron and positive ion emission of tungsten and oxidised tungsten filaments was further investigated by BECKER, *Thermionic and Adsorption Characteristics of Cæsium on Tungsten and Oxidised Tungsten* (7.150).

BECKER confirmed generally Langmuir and Kingdon's conclusions, but found that in passing abruptly from a high temperature at which the filament was free from cæsium, to a low temperature and observing the change of electron emission with time, the emission current passed through a maximum before attaining its limiting equilibrium value. This BECKER interprets as follows. At the low temperature every cæsium atom sticks to the surface and the area covered (θ) increases with time, the work function decreasing meanwhile. At the time of maximum emission θ is assumed to have the value unity, and then further increase of θ (i.e. formation of a second layer) is assumed to *increase* the work function. The product of the time taken to reach the maximum current and the rate of arrival of cæsium atoms (determined from the vapour pressure or indirectly from the positive ion current from the clean surface at high filament temperatures) was a constant independent of the vapour pressure, and gave for the number of atoms of cæsium per cm.^{2} of tungsten, $3{\cdot}7 \times 10^{14}$. This result is nearly the same as the

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number of cæsium atoms which could be packed in a single layer (a rather larger number of ions could be accommodated), confirming Becker's assumption that $\theta = 1$ when the electron emission passes through its maximum value. If the filament was initially partly covered with cæsium, the time taken at the low temperature to reach the maximum electron emission was less.

The proportional decrease in the time gave the fraction θ of the surface covered initially. BECKER found that the value of θ depended not only on the temperature of the filament and on the vapour pressure of the cæsium, but that, above the critical temperature at which positive ion emission is obtainable, θ also depended on whether electrons or positive ions were being drawn from the filament. In the latter case the previous history of the filament also affected θ. BECKER further investigated the adsorption and evaporation characteristics of the cæsium. He concludes that cæsium can evaporate either as atoms or as ions. The rate of atomic evaporation depends only on the temperature and on θ, increasing rapidly with θ for a given temperature. The ions can permanently escape from the filament only when the potential is in the right direction. The ion evaporation at a given temperature shows a maximum when θ is about 0·01. Becker's interpretation of the effects of cæsium on the thermionic emission from tungsten differs from Langmuir and Kingdon's in that BECKER finds values of θ greater than unity. He finds the work function is not linearly related to θ, and finds, moreover, that it increases for θ greater than unity. BECKER considers that at the points where the envelope line AB in Fig. (5.140*a*) touches the curves, the emission corresponds to $\theta = 1$ and the slope of the line AB then gives the work function for this case, $\phi = 1 \cdot 36$ volts. Finally, BECKER differentiates between atoms and ions in the adsorbed films.

In a later note, BECKER, *The Life History of an Adsorbed Atom of Cæsium* (7.151), reports the results of further analysis showing that the average life of a cæsium atom adsorbed on a tungsten surface equals, under equilibrium conditions, N/A, where N is the number of adsorbed atoms and A the arrival rate, both N and A being determinable by experiment. For a filament at 660° K. and an arrival rate corresponding to cæsium at 20° C., the surface is completely covered and an atom stays on it for one second. At 620° K., with a smaller arrival rate and complete covering, the life is about one minute. At times " edges " separate the filament into two regions covered to about 1 per cent. and 15 per cent. respectively, experiment showing the edge to be about 0·03 cm. wide. Atoms move from one part of the edge to the other, covering distances a million times the atomic diameter.

KINGDON, *Comparison of Thermal Electron Emission from a Pool of Cæsium and from Adsorbed Films* (7.160), measured the emission from a pool of cæsium freed as much as possible from occluded gas, and at 450° K. found $i = 5 \cdot 5 \cdot 10^{-12}$ amps./sq. in.

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Introduction of hydrogen increased the emission argon having the opposite effect. KINGDON gives the following comparative table of the emission from cæsium in different states :—

State.	Temperature of bulb.	Emission.	
		600° K.	500° K.
Cs adsorbed on oxygen film on tungsten	— 184° C.	$4{\cdot}7 \times 10^{-3}$ (obs.)	$1{\cdot}8 \times 10^{-4}$ (extrap.)
Ditto	100° C.	$1{\cdot}5 \times 10^{-5}$ (obs.)	$2{\cdot}4 \times 10^{-8}$ (extrap.)
Cs adsorbed on tungsten ..	80° C.	$5{\cdot}6 \times 10^{-7}$ (obs.)	$5{\cdot}6 \times 10^{-11}$ (extrap.)
Pool of Cs	—	—	2×10^{-11} (obs.)

A careful investigation of the emission from a completely thoriated tungsten filament carried out by DUSHMAN and EWALD, *Electron Emission from Thoriated Tungsten* (7.170), led to the following values for the thermionic constants: $\{AT^2 \exp(-b/T)\}$, $A = 3{\cdot}0$, $b = 30{,}500$. The emission from a partially thoriated filament was also measured and the results are reproduced in the accompanying table :—

$\theta = (b_\theta - b_w)/b_{Th} - b_w$.	A_θ.	b_θ.	$\log^{10} A_\theta$.
·95	1·50	31,460	·176
·83	2·08	34,150	·318
·72	3·74	36,570	·573
·56	7·76	40,070	·890
·43	10·86	42,840	1·037
·25	15·81	57,050	1·199
·18	(8·2)	48,360	·914

No simple relation could be obtained relating A_θ and b_θ. The minimum in the value of A_θ occurring near $\theta = 1$ was confirmed by measurements on other filaments.

The emission from tungsten filaments containing various oxides was examined by DUSHMAN, DENNISON and REYNOLDS, *Electron Emission and Diffusion Constants for Tungsten Filaments containing Various Oxides* (7.171), who found the emission to be similar in a general way to that of a thoriated filament. Oxide reduction and, at a lower temperature, activation occur just as for the thoriated filament, although the monatomic film obtained is less stable than for thorium owing to greater diffusion and evaporation. This made

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the determination of temperature-emission curves for the completely covered filament a difficult measurement, but the following values of the constants A and b in $i = AT^2 \exp(-b/T)$ were obtained :—

Metal.	A.	b.
Yt	7·0	31,300
La	8·0	31,500
Ce	8·0	31,500
Zr	5·0	36,500
U	3·2	33,000
Th	3·0	30,500

Values of the diffusion and evaporation constants are also given.

The problem of the cause of deactivation of thoriated filaments in valves run at high anode voltages was investigated by DAVIES and MOSS, *On the Cause of the Loss of Thermionic Activity of Thoriated Tungsten Filaments under certain Voltage Conditions* (7.180). It was confirmed that the deactivation is due to positive ion bombardment of the filament as in the experiments of LANGMUIR and KINGDON, the positive ions being produced by the bombardment of the anode by the thermionic electron stream.

KENTY and TURNER, *Surface Layers on Tungsten and the Activation of Nitrogen by Electron Impact* (7.190), investigated the emission from a tungsten filament over which was passed a stream of active nitrogen. They found the emission to be decreased, the reduction being permanent, i.e. it persisted after the stream of active nitrogen was stopped. It was found possible to restore the filament to its initial emission condition by flashing at a white heat. The effect is attributed to the formation of a layer of nitrogen on the filament.

Experimenting on the emission from a tungsten cathode in cæsium vapour, DE BRUYNE, *Layers of Cæsium and Nitrogen on Tungsten* (7.200), found that the presence of a little nitrogen in the tube gave rise under certain circumstances to a secondary peak in the emission temperature curve (see Fig. 7.200*a*). Whether the secondary peak appeared or not depended on the anode voltage, the peak being obtained only for anode voltages greater than some value between 5 and 8 volts. It was also observed that the peak took an appreciable time to develop after applying the anode voltage. DE BRUYNE suggests the explanation that when the electrons passing to the anode acquire sufficient energy (between 5 and 8 volts) they are able to generate excited or even ionised atoms or molecules of nitrogen. These deposit on the surface and hold a superimposed layer of cæsium atoms at temperatures higher than those for which the simple cæsium coating is stable. The secondary peak thus represents emission from cæsium at high temperatures.

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A new type of surface emitter was investigated by BREWER, *The Photo-electric and Thermionic Properties of Platinum-coated Glass Filaments* (7.205). Brewer's emitter consisted of a platinum filament over which was fused a thin layer of potassium glass, the outer surface being coated by sputtering with platinum. The photo-electric and thermionic emissions for different temperatures and for different electrolysis potentials between the platinum filament and the platinum coating were measured. At a given temperature a characteristic increase in the thermionic, positive ion and photo-electric emissions was observed as the electrolysis potential changed from driving potassium ions to the filament to

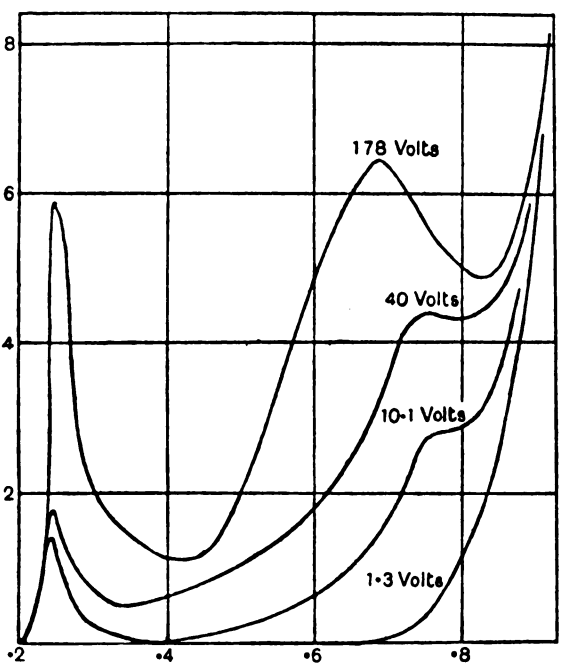

FIG. 7.200*a*.—*Filament Current in Amperes.*

driving them towards the surface. Ultra-violet light had no effect on the thermionic emission. BREWER considers that a surface solution of potassium in platinum is formed, rather than a deposit of potassium in the form of a layer on the platinum.

The linear relation between the logarithm of the emission current from tungsten covered with a surface layer of an electro-positive element, and the fraction θ of the surface covered, as commonly assumed following LANGMUIR, may be written $\log i/i_0 = \theta \log i_1/i_0$ (see 7.100). BECKER, *Thermionic Emission as a Function of the Amount of Adsorbed Material* (7.210), reports that experi-

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ments with barium layers on tungsten show that for $0 < \theta < 0{\cdot}85$

$$\log i/i_0 = 1{\cdot}1 \log i_1/i_0 \{1 - \exp(-2\theta)\}.$$

The fraction covered, θ, was determined from the time taken for the layer to be deposited. At the maximum emission θ equalled unity. BECKER suggests that an explanation may be obtained by assuming the decrease in the work function ϕ, which is chiefly responsible for the increase in log i to be due to Ba *ions* (as distinct from Ba *atoms*) adsorbed on the surface, the proportion of Ba ions to Ba atoms on the surface varying with θ from about 1 to 5 for θ small to about 1 to 10 for θ equal to unity.

EGLIN, *Thermionic Activity, Evaporation and Diffusion of Barium on Tungsten* (7.220), also examined the effect on the emission of varying the amount of barium coating on a tungsten filament. He found that with increasing thickness of coating the thermionic emission first increased and then decreased. For the maximum emission condition he obtained

$$A = 2{\cdot}5 \text{ amps./cm.}^2 \text{ deg.}^2. \quad \phi = 1{\cdot}66 \text{ volts.}$$

The same result was obtained both in the covering process and by uncovering from a thick coat of lower emission. The rate of evaporation of the barium increased rapidly with the temperature and the thickness of the coating. For the thickness corresponding to maximum emission, the rate of evaporation was estimated to be one billionth that given in the literature for barium in bulk. Up to 1,250° K. there appeared to be very little diffusion of barium into the interior of the tungsten.

The effect of surface films on the contact potential was examined by LANGMUIR and KINGDON, *Contact Potential Measurements with Adsorbed Films* (7.225). The current-potential characteristic between a test filament (cold) and a standard filament (heated to give electron emission) was determined. The test filament was then subjected to some change such as the deposition of a surface layer, and the shift of the current potential characteristic determined. This gave the difference of the contact potentials of the test filament in the initial and final states. Langmuir and Kingdon's results were as follows :—

Cs–O–W	>	W by 3·1	volts
Cs–W	>	W by 2·8	,,
Th–W	>	W by 1·46	,,
W	>	OW by 0·8	,,

A surprising result was that activated and unactivated Cs-O-W surfaces had the same contact potential, although the thermionic emissions were in the ratio $1 : 10^5$. Theoretically, the difference of contact potential ΔV should be related to the work function ϕ_1 and ϕ_2 and the constants A_1 and A_2 by the formula $\Delta V = \phi_1 - \phi_2 + \frac{T}{11{,}600} \log A_2/A_1$. The following table gives the data for ϕ and

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A for the surfaces concerned :—

Surface.	b.	ϕ.	A.	Temperature Range.
	degrees.	volts.	amps./cm.^{2} deg.^{2}.	° K.
CsOW	8,300	0·71	0·003	600– 700
ThW	30,500	2·63	3·0	1,150–1,600
W	52,600	4·52	60	1,900–2,300
OW	107,000	9·23	5×10^{11}	1,500–1,700

Computing ΔV from these data, for a temperature of 300° K., LANGMUIR and KINGDON obtain :—

	ΔV (calc.).	ΔV (observ.).
ThW–W	1·81	1·46
CsOW–W	3·55	3·1
W–OW	4·12	0·8

The agreement is described as not good. Measurements were also made of the change in ΔV and electron emission with change in temperature for a cæsium-coated filament, the amount of Cs adsorbed varying with the temperature. Fair agreement between ΔV computed and observed was obtained.

Very high values of thermionic emission were obtained by KOLLER, *Thermionic and Photo-electric Emission from Cæsium at Low Temperatures* (7.230), with films of cæsium on cæsium oxide on silver. The Richardson equation was obeyed in the range 373–443° C., above this temperature the surface deteriorating. KOLLER found $A = 9{\cdot}8 \times 10^{-2}$, $b = 8{,}700$. The cathodes used had an area of 10 sq. cm. and were formed on the bulb wall, heating being accomplished by immersing the bulb in a temperature bath. The large area made the total emission current large enough to measure at low temperatures. The photo-electric current was also measured and found to show a slight positive temperature coefficient.

Closely connected with the effect of thin films on the thermionic emission of surfaces is the shift in the photo-electric long wave limit of such surfaces. IVES and OLPIN, *Maximum Excursion of the Photo-electric Long Wave Limit of the Alkali Metals* (7.240), continuing previous work, describe experiments showing that the long wave limit for thin films of alkali metals (Na, K, Rb and Cs) moves with increasing film thickness from the blue to the red end of the spectrum, passes through a maximum excursion and then recedes. The limit at the maximum is farther out than for the alkali metal in bulk. IVES and OLPIN found that the wave length at the maximum excursion equalled that of the first line in the principal series, i.e. the resonance line, of the alkali metal concerned. Further experiments with lithium revealed a similar connexion. IVES and OLPIN draw interesting conclusions from this result, which bear on the mechanism of thermionic emission.

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NOTTINGHAM, *Influence of Accelerating Fields on the Photo-electric and Thermionic Work Function of Composite Surfaces* (7.245), repeated and confirmed Ives' and Olpin's results and investigated the analogous problem in thermionic emission. By activating a thoriated filament for various periods at the activation temperature, the emission at the lower test temperature increased first with increasing activation period, passed through a maximum, and then decreased.

The anode potential at the test temperature, which determines the applied surface field, played a fundamental part in the effect. The thickness of the thorium layer, determined by the period of activation, at which the maximum emission occurred, decreased as the anode potential increased at zero field, prolonged activation never resulting in a decrease in emission.

An optimum thickness of thorium coating was also observed by BRATTAIN, *Effect of Adsorbed Thorium on the Thermionic Emission from Tungsten* (7.247), who deposited thorium from a thoriated filament on to a tungsten ribbon. For each successive equal amount of thorium deposited the corresponding increase in the emission from the tungsten ribbon decreased until emission reached a maximum, when a decrease of emission occurred on depositing further layers, until finally a steady value was attained. Migration of thorium to the reverse side of the tungsten ribbon began to occur appreciably at 1,500° K.

According to IVES, the photo-electric emission of alkali film-covered metals varied markedly with the character of the underlying metal. EDITH MEYER, *Electron and Positive Ion Emission from Tungsten, Molybdenum and Tantalum Filaments in Potassium Vapour* (7.250), carried out experiments to test this point in the thermionic case. The electron and positive ion emissions for various pressures of potassium were obtained as functions of the filament temperature. The general form of the curves was the same as found by LANGMUIR and KINGDON, BECKER and KILLIAN (see Figs. 7.140*a*, 7.141*a*, 7.141*b*), but with this difference, that the maximum electron emission was here found to occur at the same filament temperature independently of the pressure of the potassium vapour. The emission for a given potassium vapour pressure depended markedly on the underlying metal, being greatest for tantalum and least for molybdenum. This order is not in agreement with Langmuir's suggestion that the effect of the film varies with the difference between the ionisation potential I of the alkali vapour and the work function of the underlying surface. The differences for the cases investigated are

$I_K - \phi_W = -0{\cdot}63$ volts	..	tungsten.
$I_K - \phi_{Mo} = -0{\cdot}34$,,	..	molybdenum.
$I_K - \phi_{Ta} = -0{\cdot}06$,,	..	tantalum.

The measurements indicate, it is suggested, that the thickness of the adsorbed layer is equal to several atom diameters.

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SECTION 8

OXIDE-COATED FILAMENTS

THE first investigation on the emission of electrons from heated oxides were carried out by WEHNELT. The earlier papers on the subject have already been mentioned in previous Sections under reference numbers 0.119, 0.120, 0.121, 0.140, 0.141, 0.142, 0.143, 0.160, 0.161, 0.162, 0.163, 0.164, 0.180, 0.200, 0.201, 2.150, 2.151, 3.110. The general conclusion from the earlier work was that an oxide-coated filament gives a true thermionic emission which depends only on the character of the oxide coating and is independent of the metal forming the filament.

Reviewing the results of investigations carried out in the Western Electric Company's Laboratories on the development of oxide-coated filament values for telephone repeater service, ARNOLD, *Phenomena in Oxide-coated Filament Electron Tubes* (8.110), quotes values for the average emission constants, based on temperature variation measurements on 4,310 samples. The mean values given are as follows :—

$$b = (19{\cdot}4 \text{ to } 23{\cdot}8) \times 10^3$$
$$\phi = 1{\cdot}55 \text{ to } 1{\cdot}90 \text{ volts}$$
$$A = (8 \text{ to } 24) \times 10^4$$
$$(i = AT^{\frac{1}{2}} \exp(-b/T) \text{ amps./cm.})$$

In most of the oxide cathodes used by the earlier experimenters the oxide was present as a thick layer on the filament. DAVISSON and PIDGEON, *The Emission of Electrons from Oxide-coated Filaments* (8.120), drew attention to the fact that using filaments thinly coated with oxide, an emission practically equal in amount to that given by thickly-covered filaments is obtainable.

DAVISSON and GERMER, *The Emission from Oxide-coated Filaments under Positive Ion Bombardment* (8.130), made some experiments to test the suggestion that the emission from oxide-coated filaments arises as a plentiful emission of secondary electrons from the filament, under the action of positive ion bombardment. They found this view definitely inconsistent with their results; the secondary emission was isolated and measured and found to be negligible compared with the true thermal emission.

DAVISSON and GERMER, *The Thermionic Work Functions of Oxide-coated Platinum* (8.200), made a careful study of the work function of oxide-coated platinum. The coating employed was a mixture of barium and strontium oxides, and difficulty was

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experienced owing to the fact that the state of the coating altered on changing the temperature. This resulted in an alteration of the current with the time when passing from one temperature to another. By working with the initial value of the emission it was possible to determine the temperature variation corresponding to the equilibrium state of the filament at a given temperature (1,064° K.). The resulting work function (bk/e), 1·79 volts, agreed well with that deduced from calorimetric measurements at the given temperature. For a temperature of 911° K. the equilibrium state of the filament corresponded to a work function of 1·60 volts.

An extended experimental investigation of the temperature variation of thermionic emission carried out by SPANNER, *Thermal Emission of Electrically-charged Particles* (8.300), is of special interest in connexion with oxide-coated filaments, because SPANNER concluded from his results that the electron emission of a metal and its oxide are practically identical in type and amount. Spanner's measurements were directed primarily to finding answers to the following questions :—

(1) Is there a connexion between the electron emission of a particular substance and its conductivity ?

(2) Can a relation be obtained between the thermionic emission and the electrical character (electro-positiveness or electro-negativeness) of the emitter, similar to that found for the photo-electric effect, or is there any direct relationship with the position in the periodic system of the elements ?

(3) Is the electron emission of a simple compound determinable from those of its component elements ?

SPANNER answers the first question in the affirmative and finds for the alkali earth oxides a parallel increase of thermionic emission with electrical conductivity. With respect to (2) the experimental results are found to accord satisfactorily with the following empirical formula for the constant $\phi = \frac{bk}{e}$ ($i = AT^{\frac{1}{2}} \exp(-b/T)$)

$$\frac{bk}{e} = \frac{7}{4}\frac{N^{3/2}}{Z^{\frac{1}{2}}} + 1 \text{ (volts)},$$

where N = number of valency electrons of the metal atom of the emitter and Z = its atomic number. This formula is derived from the bk/e values given in the table opposite.

The values of bk/e in the table were all obtained by applying the Richardson equation $AT^{\frac{1}{2}} \exp(-b/T)$ to the measured variation of emission with temperature. The constants A for all the oxides came out to be of the order 10^{16} e.s.u., or 3×10^6 ampere units. The fact that both metals and metallic oxides give values of bk/e, which fit in the same empirical formula, leads SPANNER to

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conclude that the electron emissions of a metal and its oxide are practically identical in type and amount. Comparative measurements with copper, nickel, tungsten and their respective oxides confirmed this conclusion. For other compounds, carbides, silicides, fluorides and sulphides, it is also stated that the emission is the same as that of the oxide. For calcium hydride, however, an increased emission was obtained (almost as much as for BaO).

RICHARDSON and BROWN, in their original investigation of the distribution of velocities among thermionic electrons (see 4.100) found that Maxwell's law was not obeyed in the case of oxide-coated filaments. Later papers (summarised in Section 4) dealing

Emitter.	Atomic Number of Metal Atom.	$bk/e = \phi$.
		volts.
BeO	4	3·45
MgO	12	3·01
CaO	20	2·40
SrO	38	2·15
BaO	56	1·85
B_2O_3	5	4·15
Al_2O_3	13	3·95
Sc_2O_3	21	3·60
Y_2O_3	39	3·10
SiO_2	14	4·8
TiO_2	22	4·0
ZrO_2	40	3·6
ThO_2	90	3·15
Ta	73	4·2
W	74	4·5
Pt	78	6·0

with this point indicate a conflict of evidence, some workers (4.190, 4.192, 4.220, 4.300) finding Maxwell's law valid for oxide cathodes, others (4.100, 4.191, 4.210) finding deviations, for the most part in the direction of an excess of high velocity electrons.

KOLLER, *Electron Emission from Oxide-coated Filaments* (8.310), in a study of the emission from platinum-iridium filaments coated with oxides of Ba and Sr by repeated dipping in a solution of the nitrates and subsequent burning in a CO_2 atmosphere, found for the emission constants ($i = AT^{\frac{1}{2}} \exp(-b/T)$), $A = 0{\cdot}00107$ amp./cm.^{2} deg.^{2}, $b = 12{,}100$ deg. KOLLER found that a trace of oxygen introduced into the tube enormously decreases the emission, that positive ion bombardment increases the emission, and that flashing at 1,600° K. deactivates the filament, which is restored by heat-

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ing at lower temperatures (900° K.). KOLLER interprets his results as showing that the emission from the oxide-coated filament is due to a film of metal formed by reduction of the oxide. Oxygen, of course, reconverts the metal to oxide and decreases the emission. Positive ion bombardment is supposed to reduce the oxide and increase the emission. Flashing at high temperature results in evaporation of the metal film, which is restored by further reduction of oxide by running at a lower temperature.

Further evidence for Koller's view that the emission from oxide-coated filaments is due to the metal formed by reduction of the oxide was obtained by GLASS, *Variation with Temperature of the Work Function of Oxide-coated Platinum* (8.311), working with commercial valves.

Papers dealing with the measurement of the cooling effect for oxide cathodes are mentioned in Section 3 (3.122, 3.123, 3.140). In one of these ROTHE, *Exit Work of Oxide Cathodes* (8.312), concludes from a study of the change in emission over long periods, that the high emission of oxide cathodes is due to metallic particles produced by chemical decomposition and deposited in the interstices of the oxide coating.

For a possible theoretical explanation of the large values of A obtained for oxide-coated cathodes see RASCHEVSKY (1.271).

The phenomena observed during the initial heating of an oxide cathode were investigated by DETELS, *The Forming Process in Oxide Cathode Tubes* (8.320). To begin with, no emission occurred; high emission then followed with the evolution of gas causing a blue glow in the tube. The gas was shown to be oxygen, proving that decomposition of the oxide was taking place. Measurements were made of the temperature emission characteristic yielding values of ϕ and A $\{i = AT^2 \exp(-e\phi/kT)\}$, two methods being employed to determine the temperature: (*a*) resistance increase of the filament; (*b*) the change in the velocity distribution of the emitted electrons (assumed Maxwellian). DETELS gives a curve showing the variation of ϕ with number of hours processing. ϕ drops from an initial value of about 7 volts to a final value of about 1·2 volts. Throughout processing the relation $\phi = \alpha + \beta \log A$ was found to hold good (α and β constants). DETELS concludes from his work that initially the area of emitting surface of the oxide coating is very large (large A), but that during processing decomposition of the oxide occurs, and the pure metal melts to form a smooth surface (small A), the emission being attributed to the pure metal. To explain why oxidation of the pure metal does not occur, DETELS assumes that an alloy is formed with the metal of the filament core.

An experimental investigation of the mechanism of electron emission from oxide cathodes (BaO, CaO, SrO) was carried out by ESPE, *Emission Mechanism of Oxide Cathodes* (8.330), who measured the variation of the emission current with (*a*) time, (*b*) anode

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potential, (*c*) temperature. ESPE summarises his conclusions as follows :—

(1) The oxide covering is split up electrolytically under the action of an initially very small emission current. Particles of the pure alkaline earth metal produced by electrolysis in the coating and subsequently diffused to the surface, must be regarded as the actual emission centres.

(2) Failure to obtain saturation currents is explained as due to the effect on the electrolysis of the anode voltage.

(3) The space charge limitation $V^3/_2$ law is well satisfied throughout. Residual gas could not, therefore, have played an essential rôle in the researches.

(4) The emission as a function of temperature may be represented for a constant chemical state by Richardson's or Dushman's equation. Changes in the chemical state result in alteration of the constant A, *b* remaining constant.

ESPE, *Exit Work of Electrons Emitted by Alkaline Earth Oxide Cathodes* (8.331), found the following values for the constant *b* :—

CaO	22,400 ± 300
SrO	16,600 ± 250
BaO	12,900 ± 250

Methods used commercially in the manufacture of oxide cathodes are described in some detail by STATZ, *Technical Preparation of Oxide Cathodes* (8.338). A method for coating small platinum electrodes with a smooth coating of oxide is given by SCHEMAEW, *Oxide-coated Filaments and some of their Properties* (8.339). SCHEMAEW found that a filament, carefully outgassed by prolonged heating *in vacuo*, gave off a great deal more gas when giving its emission current. The effects on the activity of the filament of overheating and the presence of mercury vapour (which may sometimes be favourable) are also discussed by SCHEMAEW.

In a discussion of various methods of obtaining barium oxide-coated filaments, HODSON, HARTLEY and PRATT, *Manufacture of Barium Oxide Filaments* (8.340), dealt with the following methods :—

(1) Melting barium compound on to the core wire.

(2) Application of a paste containing a barium compound.

(3) Evaporation of a solution of a barium compound on to the core wire.

(4) Deposition of barium from the vapour on to the previously oxidised tungsten core.

They found method (4) to give the most constant filament emission. Barium metal-coated filaments and barium oxide-coated filaments

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by the evaporation method showed no difference in properties, except that the metal coating on unoxidised base filaments scarcely persists long enough to obtain measurements, whereas the oxide coating remains unchanged for thousands of hours.

The view that the low values of the constant A in Richardson's equation obtained for barium oxide cathodes are to be attributed to a very small effective emitting area, consisting of patches of pure barium produced by electrolysis of the oxide, was advanced by ESPE, *Richardson Constants of Distillation Cathodes* (8.350). If only a fraction of the surface is covered with metallic barium the emission equation may be written $i = fAT^2 \exp(-b/T)$. Since Af for oxide cathodes is at most 10^{-3}, it follows, by comparison with the A values of other thin film emitters ($A = 3$ for completely thoriated tungsten), that f cannot be greater than about a thousandth. A barium emitter formed by distilling barium on to the filament would be expected to give $f = 1$. ESPE found that, using the distillation method and depositing the barium on to an oxidised tungsten filament, the constant b was practically the same as for oxide-coated filaments produced by the paste method. The constant A, however, was much increased (100 to 1,000 times greater) in accordance with the theoretical prediction. In the absence of the intermediate tungsten oxide layer the emission soon fell to the value for the tungsten core (compare 8.340). The following table given by ESPE compares his present results with other determinations of the Richardson constants :—

Investigator.	Emitter.	A.	b.	ϕ.
ESPE (1929) ..	Ba on oxidised tungsten—			
	1st activation ..	$1{\cdot}1 \times 10^{-1}$	12,600 ± 500	1·09 ± 0·04
	2nd activation	$3{\cdot}0 \times 10^{-1}$	12,800 ± 500	1·1 ± 0·04
ESPE (1927) ..	BaO paste (thin)	$3{\cdot}0 \times 10^{-4}$	11,500 ± 300	0·99 ± 0·03
KOLLER (1925)	BaO and SaO paste	$1{\cdot}07 \times 10^{-3}$	12,100	1·04
DUSHMAN and EWALD (1927)	Th on W ($\theta = 1$)	3·0	30,500	2·62
LANGMUIR, KINGDON, BECKER (1925 and 1926)	Cs on W ..	—	15,800	1·36
KINGDON (1924)	Cs on W ..	1×10^{-3}	8,300	0·71

Measurements of the cooling effect for the distillation cathode yielded a mean value 1·21 volts, i.e. 10 per cent. greater than the temperature variation value. This could be regarded as satisfactory agreement, bearing in mind the experimental errors involved.

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In MACNABB'S investigation of the formation of oxide-coated filaments by different methods, *The Production of Emission from Oxide-coated Filaments—A Process Phenomenon* (8.360), attention was paid to the relative amounts of "combined" and "uncombined" oxide in the product. MACNABB concludes that the fundamental principle underlying all methods is that the filament must undergo gaseous bombardment before it will emit, the effect of the bombardment being a reduction of higher oxides or the carbonates to the lower oxides or pure metal. The results suggest that the most effective gas for the purpose is CO_2, the most suitable coatings being those containing most "uncombined" oxide. Changing the core metal had no effect.

An appreciable increase in the emission from barium oxide-coated filaments was obtained by BEESE, *Thermionic Emission of Oxide-coated Cathodes containing a Ni-Ba Alloy Core* (8.365), by introducing a small percentage of metallic barium (approx. 0·15 per cent.) into the nickel core material. The increase was particularly marked in the case of lightly-coated filaments.

A comprehensive investigation of the processes occurring in oxide-coated filaments was carried out by BECKER, *Phenomena in Oxide-coated Filaments* (8.400). BECKER works out a theoretical explanation of the observed effects. Two types of oxide-coated filaments are distinguished: (*a*) with some or all the oxides combined chemically with the metallic core; (*b*) with all the oxides uncombined. In both types, particularly the combined type, the break-down process (glowing at 1,500° K. for several minutes and then applying a high positive potential gradient to the filament for from 2 to 3 minutes) enhances the emission. BECKER takes as his first hypothesis the suggestion previously made by KOLLER (8.310), ROTHER (8.312), ESPE (8.330), and DETELS (8.320), that the enhanced activity of oxide-coated filaments is due to metallic barium on the surface of the barium oxide. In other composite surfaces, however, change in the anode voltage does not affect the emission to the same degree as for oxide-coated filaments. Hence some further assumption must be made, and BECKER follows ROTHE, ESPE and DETELS in postulating electrolysis of the oxides. In support of this BECKER points out that oxygen is emitted and barium diffuses into the core, if space current is taken from the filament when it is being operated at high temperatures; but neither of these effects is observed if no space current is passed.

BECKER submits experimental evidence showing:—

(1) That metallic barium evaporates from a well broken-down coated filament.

(2) That oxygen evaporates if the maximum emission current is drawn from the filament and the temperature is high enough.

(3) That high activity of oxide-coated filaments is due to metallic barium on the oxide surface.

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(4) That barium diffuses readily to and from the oxide surface.
(5) That oxygen affects the activity in various ways.
(6) That oxygen diffuses readily to and from the oxide surface.
(7) That most of the current through the oxide is carried by electrons, but a small part is carried by ions, the ratio depending on the temperature, the composition of the oxide layer and the anode potential.
(8) That the complex time changes that occur can be interpreted in terms of these processes.

Becker's conclusions may be summarised as follows. The activity of a coated filament depends on the surface concentration of barium and oxygen, on the relative positions of the barium and the oxygen, and on the composition of the underlying oxide layer. Barium can be deposited on the surface (1) by distillation from an outside source, (2) by electrolysis, sending electrons *into* the filament. The barium deposited on barium oxide behaves similarly to barium on tungsten; as the amount of barium increases the logarithm of the emission current increases, rapidly at first and then more slowly, to a maximum value. Oxygen also can be deposited on the surface by introduction of oxygen gas or by electrolysis (drawing electrons out of the filament). The effect of oxygen on the activity is complex, but, in general, similar to its effect in the case of Ba on O on W. If the oxygen lies between the barium and the oxide activity is increased, but if it is superimposed on the barium the activity is reduced. Both barium and oxygen diffuse readily from the oxide to the surface and vice versa, the rate of diffusion increasing rapidly with temperature and with the concentration gradient near the surface. When current is sent into or drawn out from the oxide filament, most of it is carried by electrons, but a small proportion is conveyed by barium and oxygen ions. The proportion depends on the temperature, the composition of the oxide layer, and, in the case of drawing electrons out, on the anode potential.

RIEMANN and MURGOCI, *Thermionic Emission and Electrical Conductivity of Oxide Cathodes* (8.410), made measurements of the conductivity of oxide coatings by twisting together two coated filaments and determining the current between them in the intervals of heating, using a special commutator to switch on and off the heating and measuring currents. The conductivity was found to vary exponentially with the temperature according to the law $C = \alpha \exp(-\beta/T)$ (α and β constants). The temperature variations of the conductivity and the thermionic emission were closely related, a linear relation obtaining between the logarithms of the conductivity and the emission current. During processing the conductivity and the emission grew and decayed together. At current densities comparable to those carried in thermionic emission the conductivity current ceases to obey Ohm's law and shows signs

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of becoming saturated. The effects of the following "poisons" were also examined :—(*a*) oxygen, (*b*) discharge in CO, (*c*) discharge in H_2. Complete recovery followed successive applications of (*c*), but only a few recoveries after (*a*) or (*b*). The authors put forward a theory of the action of oxide cathodes which may be summarised briefly as follows. The coating conducts the space current electrolytically. Only the metallic ions are mobile, the oxygen ions playing no active part in the electrolysis. The whole surface of each oxide crystal is completely covered with a mobile monatomic layer of alkaline earth metal. The passage of space current is accompanied by a continual circulation of metal diffusing outwards along the surfaces of the crystals and inwards through the crystals in the form of electrolytic ions. Space current passes from the core metal to the coating, mainly in the form of thermionic electrons produced by the barium-contaminated core surface. On the basis of their theory RIEMANN and MURGOCI offer explanations for (*a*) the eventual "life-failure" of oxide cathodes, (*b*) poisoning effects, (*c*) variations in the published values of thermionic constants of oxide cathodes.

GEHRTS, *Electron Emission from Oxide Cathodes* (8.420), reviewing the evidence on the mode of formation and action of oxide cathodes, arrives at the following conclusions, which differ from those of other investigators, particularly in the substitution of thermal dissociation for electrolysis as the mechanism of production of the metallic barium :—

(1) The electron emission of oxide cathodes is a thermionic emission from metallic barium (or other alkaline earth metal), adsorbed on the surface or within the oxide layer.
(2) The metallic barium is produced by thermal dissociation of the barium oxide.
(3) During activation, the free oxygen is removed completely from the cathode.
(4) Activation is aided by a previous "forming" process, in which a reaction occurs between BaO and an easily oxidisable constituent of the core, such as iridium, rhodium or nickel.
(5) Unformed cathodes tend to be unstable.
(6) Variations in emission in operation or due to temperature variations, arise from changes in the area of the surface film of barium produced by diffusion to and fro of the metallic barium.

Evidence that the core metal has a vital function in the phenomenon of emission by oxide-coated filaments was brought forward by LOWRY, *Rôle of the Core Metal in Oxide-coated Filaments* (8.500). LOWRY made temperature-power and temperature-emission measurements on oxide-coated platinum, oxide-coated

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Konel (alloy of Ni, Co, Fe and Ti) and uncoated filaments. For the same power radiated per cm., despite the greater thermal emissivity and hence lower temperature of the coated Konel filament, the electron emission was greater than for the coated platinum filament. The coated Konel filament gave an emission at 775° C., equal to that of the coated platinum filament at 950° C. LOWRY assumes that the metallic barium is situated not at the oxide-vacuum surface, but at the core-oxide surface, electrons having to diffuse through the oxide layer. An explanation of the effect of the core metal is in this way made possible. It follows also that the closeness of packing of the oxide layer may affect the emission current, gradual decay in emission being attributable to the closing of pores in the oxide coating. An explanation of the lack of saturation for oxide-coated filaments might be obtained on the same lines.

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	0.170	0.171	0.179	0.183	0.184	0.185	
	0.186	0.195	0.196	0.204	0.213	0.241	
	0.242	0.250	0.280	1.100	1.111	1.112	
	1.115	1.121	1.150	1.161	1.204	1.220	
	1.221	2.125	2.141	2.160	2.241	4.100	
	4.101	4.102	4.110				
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ROBERTSON, F. S.	0.280						
ROSSIGER, M.	4.190						
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SACKUR, O.	1.207						
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SCHLICHTER, W.	0.230	2.120					
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SODDY, F.	0.160	0.163					
SOMMERFELD, A.	1.400						
SPANNER, H. J.	2.250	3.122	8.300				
STEAD, G.	2.180						
STOECKLE, E. R.	2.140						
TETRODE, H.	1.208						
TIERI, L.	3.160						
TING, S. L.	4.150						
TOLMAN, R. C.	1.200						
TONKS, L.	1.290	1.292					
TURNER, L. A.	7.190						
TYKOCINSKI-TYKOCINER, J.	6.130	6.131					
van VELZER, H. L.	5.230						
VIOHL, A.	3.150						
van VOORHIS, C. C.	3.130	3.131	3.169	3.170			
WAHLIN, A. B.	0.410						
WARNER, A. H.	2.281						
WATERMAN, A. T.	1.230	5.451					
WEHNELT, A.	0.119	0.120	0.162	0.180	0.181	0.187	
	0.188						
WEIGLE, J. J.	1.250	1.251	1.252				
WILSON, H. A.	0.110	0.151	1.101	1.102	1.114	1.206	
WILSON, W.	1.120	3.110					
YOUNG, A. F. A.	2.240	2.241					
ZWIKKER, C.	0.425	1.530	2.270	2.271			

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DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH

RADIO RESEARCH

SPECIAL REPORT No. 12

A THEORETICAL AND EXPERIMENTAL INVESTIGATION OF HIGH SELECTIVITY TONE-CORRECTED RECEIVING CIRCUITS

BY

F. M. COLEBROOK, B.Sc., D.I.C., A.C.G.I.

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PREFACE

IN April, 1931, the Radio Research Board of the Department of Scientific and Industrial Research appointed a Committee to examine and report on the properties of very highly selective receivers. The Committee consisted originally of the following members :—

Professor E. V. Appleton, D.Sc., F.R.S. (*Chairman*).
Mr. F. M. Colebrook, B.Sc.
Colonel A. C. Fuller, O.B.E.
Mr. E. B. Moullin, M.A.
Mr. R. A. Watson Watt.

Dr. S. Brydon was co-opted on to the Committee, which, in the final stage of its investigation, had also the advantage of consultation with Colonel A. S. Angwin of the Post Office and Mr. Noel Ashbridge of the British Broadcasting Corporation.

The Committee has confined its attention to receivers in which high selectivity is attained by the use of radio-frequency circuits of exceptionally low decrement, combined with audio-frequency tone-correction, and to answering in this connection the two following questions :—

(i) Is the performance of such receivers consistent with hitherto accepted theory ?

(ii) What are the advantages or disadvantages of such receivers in relation to the practical problem of discriminating between wanted and unwanted wireless transmissions ?

At its first meeting the Committee formulated proposals for critical experiments, which have been carried out partly by Dr. S. Brydon at the Signals Experimental Establishment, Woolwich, and chiefly by Mr. F. M. Colebrook at the National Physical Laboratory, Teddington.

The Committee's answers to the questions stated above are given in a survey of the subject which has been drawn up for the Committee by Mr. F. M. Colebrook, and which is published herewith. In this survey the experiments carried out by him on behalf of the Committee are described, and the deductions drawn from them are stated in quantitative terms. The theoretical analysis given in the survey is for the most part quite general in character and applicable to any form of radio-frequency resonance curve, but for purposes of detailed illustration and experimental verification special attention has been given to the simple case of a single radio-frequency circuit of abnormally low decrement. As this survey is necessarily very technical, it has been thought desirable to summarise, as far as is possible without recourse to mathematical formulæ, the answers given to the two questions.

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In connection with the first question, it may be stated at once that the behaviour of receiving circuits of very high selectivity combined with tone-correction (including circuits in which the high selectivity is attained by the use of quartz crystal resonators) has been found to be quite in accordance with that predicted from an application of the usual side-wave analysis. The Committee has, however, called attention to the fact that in this type of reception certain known phenomena, associated with detection, acquire a greatly enhanced importance in consequence of the high selectivity of the receivers. The significance of these phenomena in relation to high selectivity had not previously, in the opinion of the Committee, been given due weight. This matter is discussed at length in the following survey.

In considering the answer to the second question, a distinction must first be drawn between the two types of interference which are experienced in practical reception. These can be illustrated by reference to the problem of the reception of a certain wanted modulated transmission in the presence of another, unwanted, modulated transmission of neighbouring carrier-wave frequency. (Such modulation in either case may be, for example, telephonic, as in broadcast reception, or telegraphic, as in the reception of morse signals.) The first of the two types of interference has the character of the modulation of the unwanted transmission and will, therefore, be in general directly intelligible. The second type of interference arises from the interaction of the carrier wave of the wanted transmission, on the one hand, with the carrier wave and any side waves of the unwanted transmission on the other hand. This interaction is a necessary consequence of rectification, and will, therefore, be mainly associated with the detector stage. It must be emphasised, however, that some degree of rectification is likely to occur in other stages of the receiver. Interference of this second type will, when it arises from direct telephonic modulation, differ from that of the first type in being unintelligible.

Briefly, the distinction between the two types is that the first would exist, generally in greater intensity, if the wanted carrier were removed, whereas the second would disappear.

An example of both types of interference is afforded by the attempt to receive a transmission from the London Regional station on 843 kc/s., at a point where the indirect ray from the station at Mühlacker on 832 kc/s. gives an average field strength at night of 50 per cent. or more of that of the London Regional station. If a normal receiver of variable selectivity be used, it will be found that when it is adjusted for low selectivity, and tuned accurately to the London Regional transmission, the programme from the Mühlacker station will also be audible, and in general intelligible. As the selectivity is increased the volume of the intelligible interference will decrease and an unintelligible form of interference, previously masked by the other, will consequently become more noticeable,

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the two kinds being heard intermingled. As the selectivity is further increased the intelligible interference, which has been classified above as interference of the first type, will disappear and there will remain only the unintelligible interference of the second type caused by the heterodyning of the side waves of the Mühlacker station with the carrier wave of the London station.

In the above example, the steady heterodyne note due to interaction of the two carriers, which would in general be the most noticeable component of the second type of interference, has been mitigated by an increase of the carrier frequency separation beyond the usual 9 kc/s. In the particular case of the broadcast transmissions from Warsaw and the Eiffel Tower, the carrier frequencies of which are at present separated by 5 kc/s., the steady note of this frequency is very prominent.

The Committee is of opinion that, so far as concerns immunity from the first type of interference, namely, that which retains the character of the modulation of the unwanted transmission, any form of selectivity which differentiates highly in favour of the wanted carrier wave as against the unwanted carrier wave will possess very valuable properties. With either " square-law " or " linear " detectors, the enhancement of the voltage produced at the detector by the wanted carrier wave, relative to that produced by any unwanted waves, brings about a more than corresponding reduction of the received intensity of any unwanted modulation relative to that of the wanted modulation, an advantage which remains unimpaired by the subsequent tone correction which is necessarily employed with such receivers. Theoretically, the reduction of interference of this type can be effected to almost any desired degree by increasing the selectivity or sharpness of tuning of the receiver; but in practice the permissible sharpness of tuning will be limited by conditions of circuit and transmission frequency stability, and cannot be carried beyond these limits without introducing distortion. These limits are likely to be the more stringent the higher the carrier frequency. The investigations of the Committee have been confined to frequencies less than about a million cycles per second.

In considering the second type of interference, a further distinction must be drawn between the audio-frequency components of such interference which lie (*a*) within and (*b*) without the audio-frequency range required for satisfactory reproduction of the wanted modulation. The use of highly selective receivers with corresponding tone-correction does not bring about any special reduction of (*a*) except in certain particular cases where a controllable asymmetry of the resonance curve may be utilised to produce, with some little sacrifice of quality in the case of telephony, a virtual blindness over a certain narrow band of radio-frequencies on one side or the other of the wanted carrier frequency. The (*b*) interference, on the other hand, i.e., heterodyne interference with the wanted carrier

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wave giving audio-frequencies outside the range required for satisfactory reproduction of the wanted modulation, can be eliminated from this as from other types of highly selective reception, by the general method of restricting the audio-frequency response of the receiver to the range required for the wanted modulation. This cut-off may, for example, be of practical value in the reception of " tonic-train " telegraphy with fairly low modulation-frequency.

The main result of the Committee's investigation is, therefore, to demonstrate that the combination of very high selectivity with suitably designed tone-corrected and, if necessary, frequency-limited audio-frequency amplification affords a means of effectively reducing that interference which is due to the reception of the direct modulation of a transmission on a neighbouring carrier frequency, but it is clear that no immunity is afforded against those heterodyne interactions involving the wanted carrier wave which produce audio-frequency components within the range required for the wanted modulation. In fact, in relation to the reception of broadcast telephonic transmissions of the normal double side-band type, a receiver designed on these or any other known and practicable principles can only be expected to give reception entirely free from interference arising from interaction between the wanted carrier and unwanted side-waves, provided its response is limited to those audio-frequencies which are less than half the frequency-difference between the two transmissions, and provided the audio-frequency range of the modulation of each transmission is similarly limited. It follows that the minimum carrier-frequency separation between transmissions is ultimately determined by the audio-frequency range considered necessary for satisfactory reproduction of the modulation, and cannot be reduced beyond the limit so determined without involving either an increase of interference or a corresponding lowering of quality.

Finally, the Committee has called attention to the fact that the investigation has been mainly concerned with the fundamental principles underlying the operation of highly-selective tone-corrected receivers, and that no comparative tests have been carried out as between such receivers and receivers of other types. They have further pointed out to the Board that certain special problems, such as, for example, the response to modulation of an impulsive character, have not been quantitatively examined. They did not, however, consider it desirable to delay the submission of the report for the considerable time that would have been required for the completion of the investigation in such respects.

A. G. LEE
(*Chairman, Radio Research Board*).

DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH,
16, OLD QUEEN STREET,
LONDON, S.W.1.
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A THEORETICAL AND EXPERIMENTAL INVESTIGATION OF HIGH SELECTIVITY TONE-CORRECTED RECEIVING CIRCUITS

ABSTRACT

The paper describes a theoretical and experimental investigation of the behaviour of receiving circuits of abnormally high selectivity in relation to modulated continuous waves.

Part I is devoted to a comprehensive and, as far as possible, generalised presentation of the theory of the subject within the limits imposed by the assumption of quasi-stationary conditions of electrical oscillation. Particular attention is given to the variation of the intensity of the modulation-frequency output, for both square-law and linear systems of rectification, considered as a function of modulation frequency for a given carrier frequency, and as a function of carrier frequency relative to circuit frequency for a given modulation frequency. Interference characteristics are deduced from the nature of the response to simultaneous modulated-wave excitations on neighbouring frequencies, for both types of rectification. It is shown that the side-band attenuation associated with reception at very high selectivity can be exactly compensated by a suitably designed audio-frequency amplifying system, and that the selectivity of the circuit is not changed by this process.

Part II describes an experimental confirmation of the theory by a set of experiments on ordinary tuned coil receiving circuits in which the " power-factor," or ratio of resistance to inductive reactance, is reduced to a value as low as 10^{-4} by means of valve retroaction. It also gives the measured audio-frequency response curve of a tone-corrected highly selective circuit.

Part III is devoted to an examination, partly theoretical but mainly experimental, of the behaviour of circuits made highly selective by the incorporation of piezo-electric quartz crystal resonators. The form of the radio-frequency resonance curves of such circuits is described and explained, and the effect of auxiliary circuit elements on selectivity is discussed. The results of a large number of experimental measurements on such circuits are described and analysed, and it is shown that the observed behaviour is in qualitative and quantitative agreement with theory provided that due allowance is made for the effect of a certain degree of asymmetry in the resonance curve.

A summary of the most important results and conclusions arrived at is given in the final section of the paper.

1. INTRODUCTION

Scope of the Investigation

THE selective reception of amplitude modulated electric waves is obviously a subject of major practical importance, since the selectivity of receivers is one of the limiting factors in the allocation of wave bands for communication purposes.

It is impossible to transmit intelligence by means of an electric wave of single frequency. For communication purposes it is

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necessary that the time-variation of the wave shall be modified from its initial simple harmonic form, and an inevitable consequence of any such modification is the distribution of the energy of the wave over a frequency band of finite width, with, in general, local concentrations or " peaks " of energy at frequencies adjacent to the nominal frequency of the transmission. Some of the wider aspects of this matter have been discussed by Dr. R. L. Smith-Rose and the present writer.[1]*

It follows that a receiver of infinite selectivity, i.e., one capable of response to a single frequency only, would be unable to receive any form of intelligence transmitted by means of an electric wave. It is necessary that the receiver shall be capable of response over a finite range of frequencies, but the response need not be, and in general will not be, uniform over the range of frequencies covered by the transmission. The higher the selectivity of the receiver, the smaller will be the response to the adjacent lateral frequencies. The most familiar example of this effect is the so-called " side-band cutting " of speech or music modulated broadcast transmissions, resulting in a considerable modification of the received as compared with the transmitted modulation, the modification being in the direction of an attenuation of the higher modulation-frequencies.

It is known, however, that this effect can be eliminated by a suitable manipulation of the frequency characteristic of the audio-amplification stages of a receiver. Reception of this type has recently been brought into prominence by the introduction of receivers in which use is made of the exceedingly high selectivity of a quartz crystal operating piezo-electrically.

The following paper describes an investigation, both theoretical and experimental, of the behaviour of receiving circuits of abnormally high selectivity obtained by the use of quartz crystals and by the use of high degrees of retroaction in ordinary valve circuits. The object of the investigation was to obtain exact information about the behaviour of such circuits, particularly in relation to heterodyne interference, in order to determine whether they could be expected to give any greater immunity from interference than would be expected on the basis of existing theory.

Part I of the paper gives a summarised presentation of the theory of the highly selective reception of radio transmissions based on the accepted fundamental assumptions, and limited to quasi-stationary conditions.

Part II describes a series of experiments with receiving circuits made highly selective by means of valve retroaction, the experiments being designed to afford a quantitative test of the theory as applied to circuits of abnormally high selectivity.

Part III describes similar experiments with receiving circuits embodying quartz crystals.

* These numbers refer to the bibliography on p. 69.

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PART I

GENERALISED THEORY OF SELECTIVE RECEPTION

2. Elements of the Theory

The elements involved in the theory of selective reception are (*a*) the electric field ; (*b*) the receiver. For present purposes (*b*) will be considered in its simplest form as consisting essentially of three main parts :—

(1) Tunable radio-frequency circuits ;
(2) A detector ;
(3) Audio-frequency or direct-current circuits, together with telephones or other signal-recording apparatus.

Supersonic heterodyne or multiple frequency-change receivers are not thereby excluded from consideration, since their operation is based on the same fundamental principles as the simpler forms.

Brief mention will be made of the most important features of the above elements in relation to selective reception. Their combined action will be illustrated by consideration of the reception of the simplest types of excitation, i.e., (*a*) heterodyne ; (*b*) pure tone amplitude modulated wave ; and finally the problem of the selective reception of a wanted transmission in the presence of interfering unwanted transmissions will be considered in terms of certain simple forms of simultaneous excitation.

3. The Electric Field

At any given instant the radio-frequency electric field exciting a receiving system will be exceedingly complex in character, containing components derived from all the transmitters then in operation. There is, however, no evidence of any interaction in the medium between these simultaneous excitations, which can, therefore, be considered separately and resolved into their original simple harmonic components typified by an intensity

$$e = \mathrm{E} \sin \omega t \quad \ldots\ldots\ldots \quad 1$$

(*a*) *Simple heterodyne excitation*

The familiar heterodyne combination

$$e = e_1 + e_2 = \mathrm{E}_1 \sin \omega_1 t + \mathrm{E}_2 \sin \omega_2 t \quad \ldots \quad 2$$

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has been discussed at length by the present writer[14]. For certain purposes it is convenient to synthesise the combination into a form analogous to that of an amplitude modulated continuous wave, as follows :—

$$e = \{E_1^2 + E_2^2 + 2E_1E_2\cos(\omega_1 - \omega_2)t\}^{\frac{1}{2}} \sin\left(\frac{\omega_1 + \omega_2}{2}t + \alpha\right) \quad 3$$

where

$$\tan\alpha = \frac{E_1 - E_2}{E_1 + E_2}\tan\frac{\omega_1 - \omega_2}{2}t \quad . \quad . \quad . \quad . \quad . \quad . \quad 4$$

The square root expression gives the shape of the envelope of the familiar interference pattern obtained by the graphical addition of the component sine waves. It is very approximately sinoidal if one amplitude is very small compared with the other, but differs appreciably from a sine wave in general. The fundamental component of the wave form of the envelope is, of course, of the " beat " frequency $(\omega_1 - \omega_2)/2\pi$.

To prevent a possible misinterpretation of this synthesis, it may be pointed out that although it seems to indicate a modulated radio-frequency wave with a carrier angular frequency, or pulsatance, which is the mean of the component frequencies, it would not therefore be correct to assume, by analogy with a single frequency excitation, that tuning a receiver to this mean frequency would give the maximum response. A mathematical synthesis which is convenient for some descriptive purposes does not alter the fact that the excitation actually consists of two separate and distinct frequency components, and the calculation of the response of a tuned circuit is based on this actual physical character of the excitation.

The simple heterodyne interference pattern is of fundamental importance in radio communication, in as much as any amplitude modulation, or, more generally, any intensity variation of the electric field, can be depicted as an interference pattern of pure sine wave components.

(*b*) *Amplitude modulated continuous wave*

The heterodyne interference pattern arises from the combination of sine wave components which in general originate in different sources. The amplitude modulation of a continuous wave is, in a sense, the reversal of this process, the interference pattern (or amplitude variation) being imposed on a single source, e.g., by a microphone or morse key, and giving rise, as an inevitable consequence, to component pure sine waves symmetrically disposed in pairs about a mean or carrier frequency component.

Amplitude modulation will in general be of a complex nature, but for analytical purposes a single pure tone modulation will be considered, for whatever the character of the modulation it can be represented either as the sum of a series of such elements (for

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a periodic modulation) or as an integral with an integrand of this character (for an impulsive modulation). Such a pure tone modulated wave will be represented by an intensity

$$e = \mathrm{E}\,(1 + \mathrm{M}\cos pt)\sin \omega t \quad \ldots\ldots \quad 5$$

The right hand side can be replaced by the well-known identical expression in terms of three single frequency components, i.e.,

$$e = \mathrm{E}\sin \omega t + \frac{\mathrm{ME}}{2}\sin(\omega - p)t + \frac{\mathrm{ME}}{2}\sin(\omega + p)t^{*} \quad 6$$

$$= e_0 + e_1 + e_2 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad 7$$

In what follows e_0, e_1 and e_2 will be called the carrier wave and the lower and upper side waves respectively.

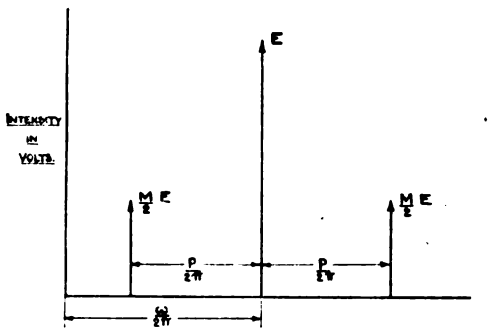

FIG. 1*a*.

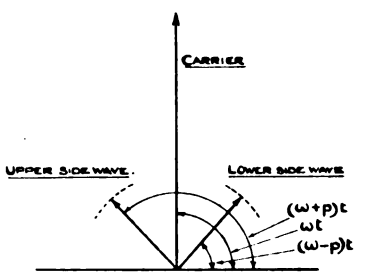

FIG. 1*b*.

The spectrum of the modulated wave is obviously as illustrated in Fig. 1*a*. The trinity of waves can also be represented by the rotating vector diagram of Fig. 1*b*. If the centre vector is thought of as stationary, the side vectors must be thought of as rotating in opposite directions, with angular velocity p. This diagram is familiar but is included here in order to emphasise

* This transformation is commonly referred to as the " side-band theory," the word " theory " being used in the sense of " hypothesis." The phrase is unfortunate, since there is no " side-band theory " in this sense. The recent controversy on the subject was a consequence of this misuse of the word " theory." (See Bibliography No. 6.)

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certain features of the modulated wave which will be found to be of importance in connection with selective reception. A little consideration will show that the phase and magnitude symmetry, illustrated in Fig. 1*b*, is an essential condition for the identity of the three wave and pure tone modulated wave systems. Thus a phase change of the carrier will merely rotate the diagram as a whole through an angle equal to the phase change, while a phase change of the modulation will turn the side vectors in opposite directions through an angle equal to the phase change, the symmetry being maintained in each case.

It will be shown in a later section (Section 6) that the effect of any departure from symmetry, caused, for example, by reception in an asymmetrically resonant circuit, will be to introduce small harmonic components into the original pure tone modulation. It appears, therefore, that whatever the nature of the reception process, it should, as far as possible, maintain unimpaired the phase and amplitude symmetry of the original spectrum of the excitation.

4. The Receiving Circuit

(*a*) *Radio-frequency elements*

The function of the tunable radio-frequency circuits of the receiver is to transform the electro-motive forces, induced by the exciting electric field, into corresponding potential differences for the operation of the detector. However elaborate the circuits may be, it can be assumed that a field intensely represented by the vector **e** (of magnitude E sin ωt) gives rise to a potential difference **v** at the terminals of the rectifier, where

$$\frac{\mathbf{v}}{\mathbf{e}} = y = \mathrm{Y}\varepsilon^{j\phi} \qquad \ldots\ldots\ldots 8$$

The complex operator y may be described as a generalised transformation ratio which will be a function of the circuit constants (including adjustable tuning components) and of frequency. The curve showing the variation of Y with frequency may be similarly described as the generalised resonance curve of the radio-frequency elements of the circuit considered as a whole. The curve will in general be approximately symmetrical about some angular frequency ω_0 (not necessarily a resonant frequency in the usual simple sense) and, in highly selective receivers, will fall steeply on either side of a narrow band of frequencies.

The simplest type of receiving circuit to take for purposes of illustration is a frame aerial, directly excited by the electric field. It will be assumed to be of inductance L and resistance R, tuned by a capacity C. The e.m.f. induced in the frame aerial will be proportional to **e** (the field intensity) and to ω (the angular frequency). If the rectifier is operated by the potential difference across the

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tuning condenser, it is easy to show that, neglecting certain constants which play no part in the phenomena to be considered,

$$y = \frac{\omega/\omega_o}{(\omega^2 LC - 1) - j\omega CR} \quad \ldots\ldots\ldots 9$$

the resonant frequency being

$$\omega_o = 1/\sqrt{LC} \text{ exactly} \quad \ldots\ldots\ldots 10$$

If we put

$$\frac{\omega - \omega_o}{\omega} = \frac{\delta\omega}{\omega} = \alpha \quad \ldots\ldots\ldots 11$$

and

$$\omega_o CR = \frac{R}{\omega_o L} = P \text{ (power factor of coil)} \quad \ldots\ldots 12$$

$$y = \frac{1}{2\alpha(1 - \alpha/2) - jP} \text{ approx. if } \alpha^2 << 1 \quad \ldots\ldots 13$$

i.e.,

$$Y = \frac{1}{\sqrt{4\alpha^2(1 - \alpha) + P^2}} \quad \ldots\ldots\ldots 14$$

and

$$\tan\phi = \frac{P}{2\alpha}(1 + \alpha/2) \quad \ldots\ldots\ldots 15$$

It will be shown in Parts II and III that in a highly retroactive, or a quartz crystal circuit, P can be made to be of the order 10^{-4}, and this figure will be taken as representative of very low power-factor (or low decrement) circuits in general.

It will appear later that α will generally be small compared with 1. In all such cases we can use the simpler formulae:

$$Y = \frac{1}{\sqrt{4\alpha^2 + P^2}} \quad \ldots\ldots\ldots 16$$

$$\tan\phi = \frac{P}{2\alpha} \quad \ldots\ldots\ldots 17$$

These indicate a desirable symmetry of resonance.

The actual resonance curve of the circuit for $P = 10^{-4}$ and $\omega_o = 2\pi \times 1{,}000$ kc/s. is shown in Fig. 2*a*, in order to give a quantitative idea of the behaviour of such a circuit. Fig. 2*b* gives the circle diagram form of the same resonance curve. It can be shown that the equation of the circle diagram is

$$y = \frac{j}{2P}(1 + \varepsilon^{2j\theta}) \quad \ldots\ldots\ldots 18$$

where

$$\tan\theta = 2\alpha/P \quad \ldots\ldots\ldots 19$$

It should be noted that if Y_m is the maximum or resonance value and Y the value corresponding to a fractional frequency displacement

$$\pm\sqrt{(Y_m^2/Y^2) - 1} = (2/P)\alpha \quad \ldots\ldots 20$$

which is of the linear form

$$y = (2/P)\,x \quad \ldots\ldots\ldots 21$$

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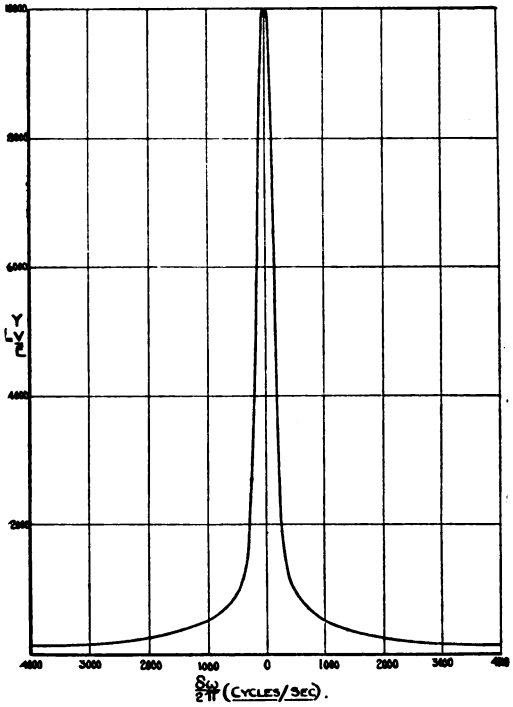

FIG. 2*a*.

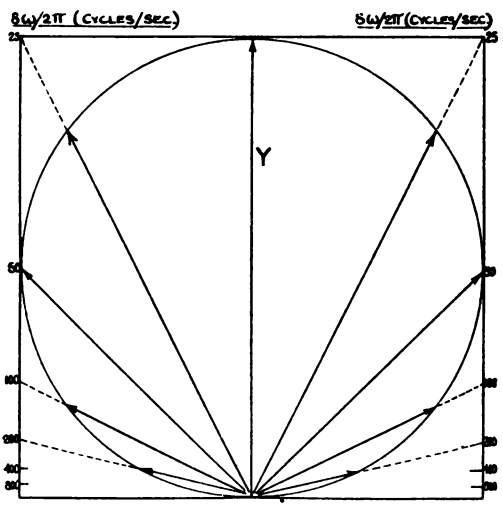

FIG. 2*b*.

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This provides a means of determining whether an experimental resonance curve is of the above simple type. The process is illustrated in section (18A) in relation to a quartz crystal resonance curve.

A simple resonance of the above type may, for convenience, be referred to as a " circular " resonance, the term being derived from the corresponding circular locus of the generalised transformation ratio y.

(*b*) *The rectification process*

A general discussion of the rectification process is obviously outside the scope of the present paper. For immediate purposes, two types of rectification will be considered as representing the extremes of all existing forms. The first is square law or small amplitude rectification, and will be specified by the assumption that the rectified output is proportional to the square of the amplitude of the input radio-frequency potential difference. The second is the so-called " linear " or " perfect " type of rectification, and is specified by the assumption that the rectified output is proportional to the mean value of alternate half-cycles of the input radio-frequency potential difference. This implies that the response to a modulated wave will be of the same wave form as the envelope of the input radio-frequency potential difference.

For simplicity, the constants of proportionality will be assumed to be unity and to be independent of modulation frequency unless otherwise stated.

(*c*) *Audio-frequency and output stages of the receiver*

This subject is only mentioned at this point to complete the descriptive scheme of Section 2, and need not be considered in any detail. It will be sufficient to point out that hitherto uniformity of response with respect to audio-frequency has been regarded as a desirable feature, at least for telephony reception. It will be shown in a later section (p. 14), however, that in reception at high selectivity a rising frequency characteristic is necessary in order to correct for the effect of the high selectivity on the audio-frequency amplitude characteristic of the modulation.

Having thus discussed the essential features of the excitation and the receiving circuit, the actual process of reception, in which each of the above elements is involved, can now be considered.

5. Heterodyne Reception

The generalised receiving circuit described in section 4 (*a*) will transform the incident wave intensities

$$e_1 = E_1 \sin \omega_1 t \quad \ldots\ldots\ldots\ldots .22$$

and

$$e_2 = E_2 \sin \omega_2 t \quad \ldots\ldots\ldots\ldots .23$$

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into corresponding rectifier potential differences

$$v_1 = Y_1E_1 \sin(\omega_1 t + \phi_1) \quad \ldots\ldots\ldots .24$$

$$v_2 = Y_2E_2 \sin(\omega_2 t + \phi_2) \quad \ldots\ldots\ldots .25$$

The amplitude of the combination, synthesised as shown in section 3 (*a*) is

$$V = \sqrt{\{Y_1^2E_1^2 + Y_2^2E_2^2 + 2Y_1Y_2E_1E_2 \cos(bt + \phi_1 - \phi_2)\}} \quad .26$$

where

$$b = \omega_1 - \omega_2 \quad \ldots\ldots\ldots .27$$

(*a*) *Square law rectification*

Assuming the rectified response to be proportional to V^2, the amplitude of the beat frequency output is

$$E_1E_2 \cdot Y_1Y_2 \quad \ldots\ldots\ldots .28$$

Considered as a function of frequency it can be shown, and is, in fact, fairly obvious by inspection, that the quantity Y_1Y_2 will in general have two maximum values corresponding very nearly (though not quite exactly) to the separate maxima of Y_1 and Y_2, which again correspond to circuit resonance with ω_1 and ω_2 (i.e., $\omega_1 = \omega_o$ and $\omega_2 = \omega_o$). The more selective the circuit, the sharper and more distinct will be the peaks of the rectified response and the more nearly will they correspond to resonance with the component excitation. Moreover, the symmetrical form of the expression shows that if the resonance curve is symmetrical, i.e., if, using functional notation, $Y(\omega_o + b) = Y(\omega_o - b)$, then the two peak values will be equal in magnitude. *Thus, in the reception of a heterodyne combination on a highly selective receiver with square law rectification there will be two distinct tuning points, giving equal response, and corresponding very approximately to circuit resonance with either component.* Experiment confirms this conclusion.

(*b*) *Linear rectification*

The beat frequency response will be the fundamental component of the Fourier analysis of the square root expression. In general, a binomial expansion of this square root will be permissible (assuming the third term to be small compared with the sum of the first two), giving as an approximation to the amplitude of the beat frequency component

$$E_1E_2 \cdot \frac{Y_1Y_2}{\sqrt{Y_1^2E_1^2 + Y_2^2E_2^2}} \quad \ldots\ldots\ldots .29$$

As in the square law case, this expression has two peak values as a function of frequency (or of circuit tuning) but the peak values will be unequal in magnitude unless $E_1 = E_2$. In fact, somewhat paradoxically, the larger response will be obtained when the circuit is tuned to the smaller of the two excitations. This is most readily seen by reference to a numerical case. Assume the simple circuit

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described in Section 4 (*a*), with $\omega_o = 2\pi \times 1{,}000$ kc/s. and $b = \omega_1 - \omega_2 = 1{,}000$ c/s. When the circuit is tuned to e_1, $Y_1 = Y(\omega_o) = 10^4$ and $Y_2 = Y(\omega_o - b) = 5 \times 10^2$. The response is therefore

$$\frac{5 \times 10^6 E_1 E_2}{(10^8 E_1^2 + 25 \times 10^4 E_2^2)^{\frac{1}{2}}} \simeq \frac{5 \times 10^6 E_1 E_2}{10^4 E_1} = 500 E_2$$

When tuned to e_2 the response is similarly

$$\frac{5 \times 10^6 E_1 E_2}{(25 \times 10^4 E_1^2 + 10^8 E_2^2)^{\frac{1}{2}}} \simeq \frac{5 \times 10^6 E_1 E_2}{10^4 E_2} = 500 E_1$$

Thus, in the reception of a heterodyne combination on a highly selective receiver with linear rectification, there will be two tuning points corresponding very approximately to circuit resonance with either component. The peak responses are very approximately in the ratio to the amplitudes of the component excitations, the larger being obtained when the circuit is tuned to the smaller component.

This apparently paradoxical result, for which experimental confirmation is given in Part II, is of some interest in connection with the reception of broadcast transmissions of very unequal intensities on neighbouring wave-lengths. With linear rectification and a highly selective circuit, the intensity of the carrier beat frequency note will be greater when the weaker station is being received. With square law rectification, the beat frequency intensity will be the same whichever station is being received. It is shown in section 8, however, that if the interference be expressed as a ratio of the intensity of the beat frequency to that of the desired modulation, the interference with linear rectification is about half that with square law rectification.

6. Reception of a Modulated Continuous Wave

Certain features of the behaviour of a tuned selective receiving circuit and a rectifier in relation to a modulated continuous wave have been considered by G. Joos and J. Zenneck[2], by R. V. L. Hartley[3], by E. B. Moullin[4], and by the present writer[5]. The following section gives a comprehensive and generalised formulation of the subject, limited, however, to the quasi-stationary condition.

Referring to sections 3 (*b*) and 4 (*a*), the three sine wave components e_o, e_1 and e_2 of the modulated wave

$$e = E(1 + M \cos pt) \sin \omega t \quad \ldots\ldots\ldots .30$$

will give rise to three corresponding rectifier potential differences

$$v_o = E Y_o \sin(\omega t + \phi_o) \quad \ldots\ldots\ldots .31$$

$$v_1 = E \frac{M}{2} Y_1 \sin\{(\omega - p) t + \phi_1\} \quad \ldots\ldots .32$$

$$v_2 = E \frac{M}{2} Y_2 \sin\{(\omega + p) t + \phi_2\} \quad \ldots\ldots .33$$

where Y_o, Y_1 and Y_2 are the generalised transformation ratios corresponding to angular frequencies ω, $\omega - p$ and $\omega + p$. It is

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not assumed at this stage that the circuit is tuned to either of the three components.

The total rectifier potential difference can be synthesised into the form of a modulated wave, but not, in general, a pure tone amplitude modulated wave. Thus

$$v_c + v_1 + v_2 = v = V \sin(\omega t + \theta) \quad \ldots\ldots .34$$

where both V and θ are periodic functions of t with a fundamental angular frequency p.

The periodic variation of θ with t is of the nature of a phase modulation. Physically, it means that the individual cycles of the radio frequency oscillation are not exactly sine waves, but provided the modulation frequency is considerably lower than the radio frequency, the deviation of individual cycles from true sine wave shape will be of negligible effect in considering the action of the rectifier. It will therefore be unnecessary to take into account in the following sections, and attention will be confined to the amplitude variation at modulation frequency.

By expansion of equations 31, 32 and 33 in the form $A \sin \omega t + B \cos \omega t$, etc., and the combination of such terms in the form $\{(\Sigma A)^2 + (\Sigma B)^2\}^{\frac{1}{2}}$, it can be shown that

$$\frac{V^2}{E^2} = \left(Y_c^2 + \frac{M^2 Y_1^2}{4} + \frac{M^2 Y_2^2}{4}\right) + M Y_c \{Y_1 \cos(pt - \phi_1 + \phi_c) + Y_2 \cos(pt - \phi_c + \phi_2)\} + \frac{M^2 Y_1 Y_2}{2} \cos(2pt - \phi_1 + \phi_2) \quad 35$$

Thus V is of the form

$$V = \{A + B \cos(pt + \beta) + C \cos(2pt + \gamma)\}^{\frac{1}{2}} E \quad \ldots .36$$

which, in the general case, represents a complete Fourier series with fundamental angular frequency p, the series containing a constant term and all the harmonics.

This result, which, for a general case, implies some small degree of harmonic distortion of the modulation wave form, will hold in any case in which, either by some degree of mistuning of the circuit, or, more usually, by some degree of asymmetry in the resonance curve of the receiving circuit, the side wave components v_1 and v_2 of v, the rectifier potential difference, are unequal in amplitude or unsymmetrical in phase difference with respect to the carrier wave potential difference. Thus, as Moullin has pointed out[4], a tuned circuit alone, irrespective of any non-linear rectifying characteristic, may be capable of producing harmonic distortion in the modulation. In circuits of high selectivity, however, the relative intensity of the double frequency term will be negligibly small, since Y_1 and Y_2 will be very small compared with Y_c.

(*a*) *Circuit tuned to resonance with the carrier frequency*

Assuming a symmetrical resonance curve, with the carrier at the centre frequency,

$$Y_c = Y(\omega_o) \quad \ldots\ldots\ldots\ldots\ldots .37$$

$$Y_1 = Y_2 = Y(\omega_o \pm p) = Y_s \quad \ldots\ldots\ldots .38$$

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and

$$\phi_c = \frac{\phi_1 + \phi_2}{2} \quad \ldots\ldots\ldots\ldots\ldots .39$$

With these conditions (which represent the normal " tuning in " of the modulated wave) it is easily shown that

$$\frac{V}{E} = Y_c + M\,Y_s \cos\left(pt - \frac{\phi_1 - \phi_2}{2}\right) \quad \ldots\ldots .40$$

$$= Y_c\left\{1 + M\,\frac{Y_s}{Y_c}\cos\left(pt - \frac{\phi_1 - \phi_2}{2}\right)\right\} \quad \ldots .41$$

Thus the original phase and amplitude symmetry is maintained, and the rectifier potential difference has the form of a true modulated wave, but the modulation percentage is reduced from its original value in the ratio of Y_s to Y_c. The extent of this reduction for any given modulation frequency can therefore be determined directly by inspection of the overall radio-frequency resonance curve of the receiver.

It is also apparent that with either type of rectification, the modulation frequency output will be proportional to Y_s (i.e., in functional notation, to $Y(\omega_c \pm p)$). Its variation as a function of p (i.e., of modulation frequency) will therefore be of the same shape as the variation of Y with ω, i.e., will be of the same shape as the radio-frequency resonance curve of the receiver. This may be of practical use as a means of determining the shape of the resonance curve (assuming this to be symmetrical).

It follows further that if a uniform overall modulation frequency response is to be obtained it will be necessary that the audio-frequency amplifier shall have a frequency characteristic of the shape 1/Y. This point will be considered in fuller detail in a later section.

The description up to this point is quite general, but certain additional deductions can be made in relation to the typical single circuit resonance specified in section 4 (*a*). For this circuit

$$Y_c = \frac{1}{P}\ (\text{reciprocal of " power factor "}) \quad \ldots\ .42$$

and

$$Y_s = \frac{1}{\sqrt{4\ \alpha^2 + P^2}} \quad \ldots\ldots\ldots .43$$

$$\simeq \frac{1}{2\ (1 + P^2/8\ \alpha^2)} \quad \ldots\ldots\ldots .44$$

where

$$\alpha = \frac{p}{\omega} = \frac{p}{\omega_c} \quad \ldots\ldots\ldots .45$$

The quantity α, or the ratio of the modulation frequency to the carrier frequency, is clearly of greater significance than p or α alone. Moullin has used the name " fractional modulation frequency " for this quantity. It is also apparent that the ratio of P to α (i.e., circuit power factor to fractional modulation frequency) is of greater significance than either P or α.

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Definition of "high selectivity."—The term "high selectivity" has been used up to this point in a purely qualitative sense. The above analysis affords a means of attaching a more definite significance to this term, at least in relation to the simple or "circular" resonance curve. It is proposed as a suitable criterion of "high selectivity" that $P^2/8\,\alpha^2$ shall be negligible compared with unity at the lowest modulation frequency involved. This criterion emphasises the fact that both the circuit constants and the frequencies of operation are equally significant in the term.

The quantity $P^2/8\,\alpha^2$ will be less than 10 per cent. of unity if the power factor is less than one-sixth of the lowest value of the fractional modulation frequency. Thus the circuit of section 4 (*a*) could be described as of high selectivity in relation to all modulation frequencies greater than about 150 c/s. on a carrier of 1,000 kc/s.

With the above interpretation of high selectivity it will be permissible to use the simpler formula

$$Y_s = \frac{1}{2\alpha}\,;\ \alpha = \frac{p}{\omega} = \frac{p}{\omega_0} \qquad \ldots\ldots\ldots .46$$

Since with either type of rectification the modulation frequency output is proportional to Y_s (i.e., inversely as the modulation frequency) it is clear that the appropriate tone correction will be given by an audio-frequency amplifier with a characteristic directly proportional to frequency.

With linear rectification it follows from equations 40 and 46 that the output has an amplitude $M/2\,\alpha$. This is independent of P (i.e., of the power factor of the receiving circuit). Thus, as Moullin has pointed out[4], with linear rectification there will be a limit beyond which the further lowering of the circuit resistance (e.g., by retroaction) will not produce any corresponding increase in modulation frequency response. It also follows as an even more important consequence, considered more fully in the section dealing with interference, that once this limit has been reached, further reduction of circuit power factor will not give any corresponding increase in circuit selectivity alone, i.e., considered in relation to separate wanted and unwanted transmissions.

With square law rectification, it is easily seen (by squaring the expression for V) that the fundamental modulation frequency output is of amplitude $M\,Y_0\,Y_s\,E^2$, i.e., $E^2M/2\,\alpha\,P$. This will increase as P is decreased (e.g., by retroaction) without theoretical limit (see, however, the limitation referred to in section 12). There is, in addition, with square law rectification, a term of double modulation frequency of amplitude $M^2Y_s^2/4$. The ratio of this extraneous double frequency term to the fundamental is $M\,Y_s/4Y_c$. The production of this double frequency term has hitherto been regarded as a serious objection to the use of square law rectification for the reproduction of speech or music modulation, an objection which outweighs the fact that it may be necessary to use radio-frequency amplification in order to get signal amplitudes large

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enough for linear rectification. It is important to notice, however, that with reception at high selectivity the objection is very much less serious, owing to the great reduction in effective modulation percentage at the rectifier, due to the high selectivity. It is true that as a result of the post-rectifier tone-correction, the double frequency term is amplified twice as much as the fundamental, but the ratio, taking this into account, is still only $MY_s/2Y_c$, and for all except the lowest audible frequencies this will be very small even with 100 per cent. modulation. Thus for the circuit already specified, the ratio is $P/4\alpha$ with 100 per cent. modulation in the incident wave. Taking P as 10^{-4} and α as 5×10^{-4} (i.e., a modulation frequency of 500 cycles per second with a carrier frequency of 1,000 kc/sec.), the harmonic ratio is only 5 per cent. and it will be correspondingly less than this for all frequencies above 500.

Thus, provided a tone-corrected receiving circuit can be made sufficiently selective in relation to the given fractional modulation frequency, there would seem to be no reason why square law rectification should not be considered permissible for the reception of speech or music modulation. This would imply less radio-frequency amplification and a corresponding freedom from the cross modulation interference associated with partial rectification in the radio-frequency stages.

(*b*) *Variation of response with circuit tuning*

Some practical interest attaches to the shape of the modulation frequency output variation as a function of carrier frequency or circuit tuning. In certain schemes of overall receiver tests, for example, the specification of selectivity is based on the shape of this curve for some arbitrary constant modulation frequency. It will be shown that such a specification is quite unsuitable for receivers of high selectivity.

We are concerned with the variation with carrier frequency ($\omega/2\pi$) of the audio-frequency components of the quantity

$$\frac{V^2}{E^2} = Y_c^2 + \frac{M^2(Y_1^2 + Y_2^2)}{4} + M\,Y_c\{Y_1 \cos(pt - \phi_1 + \phi_c) + Y_2 \cos(pt - \phi_c + \phi_2)\} + \frac{M^2\,Y_1\,Y_2}{2}\cos(2\,pt - \phi_1 + \phi_2) \quad . \quad . \quad .47$$

where Y_c, Y_1 and Y_2 are the generalised transformation ratios corresponding to the carrier, lower, and upper side angular frequencies ω, $\omega - p$, and $\omega + p$.

The analysis of the variation is exceedingly complex, and in the present paper only certain critical features will be considered, on the general lines indicated by Moullin[4].

(A fuller analysis by the present writer will be found in the paper listed as No. 5 in the bibliography.)

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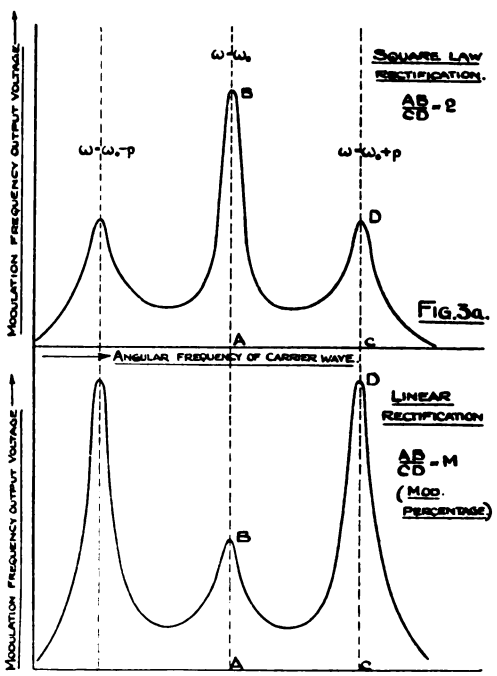
The centre peak value has already been determined, and for the simple typical resonant circuit of section 4 (*a*), is given by

$$\frac{V}{E} = \frac{1}{P}\left\{1 + \frac{MP}{2\alpha}\cos\left(pt - \frac{\phi_1 - \phi_2}{2}\right)\right\} \quad \ldots .48$$

In addition to the centre peak there will be two side peaks corresponding very approximately (though not quite exactly) to circuit resonance with the side frequencies. The exact expression for the side peaks is very cumbersome, but a close approximation for the simple circuit is given by

$$\frac{V}{E} = \frac{1}{P}\left\{\frac{M}{2} + \frac{P}{2\alpha}\sin pt + \frac{MP}{8\alpha}\sin 2\,pt\right\} \quad \ldots .49$$

Thus with square law rectification, the centre peak of the resonance curve will be proportional to $M/\alpha P$ (neglecting the relatively small double-frequency term) and the side peaks will be of half this height, as shown in Fig. 3*a*.



FIGS. 3*a and* 3*b*.

With linear rectification, on the other hand, the centre peak will be of height proportional to $M/2\alpha$ and the side peaks of height proportional to $1/2\alpha$, so that if M is less than 100 per cent. the side peaks will actually be higher than the centre peak, as shown in Fig. 3*b*. This result had previously been found experimentally by the author[5] but no explanation was then found for it. The result is, obviously, very similar to that already described in relation to heterodyne reception. In fact, when the circuit is tuned to

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either side wave, the contribution of the further side wave is negligible, due allowance being made for the larger magnitude of the carrier and its closer proximity to the resonant frequency. The case is, therefore, very nearly the same as that in which the circuit is tuned to the smaller component of a heterodyne combination.

With circuits of high selectivity, curves such as those of Figs. 3*a* and 3*b* are obtained for the variation of the modulation frequency response at modulation frequencies as low as 250 cycles per second. Their shape is a complicated function of the circuit selectivity, the modulation frequency, and the type of rectification, and such curves are therefore quite unsuitable for the specification of circuit selectivity.

It may be well to point out that the centre tuning point is the only one of the three which gives a critical value for all modulation frequencies, and is, therefore, the point to which a receiver would be adjusted by ear in the case of a complex modulation.

An apparently anomalous case. Two peaks only.—It is possibly a familiar observation that in tuning a retroactive detector to, say, the tuning note of one of the London broadcasting stations, there appear to be two peaks of modulation-frequency intensity. One could understand one or three peaks, but two peaks seem to present an anomaly.

The probable explanation is as follows. Assuming effectively linear rectification, the variation should be as in Fig. 3*b*, if the " resolving power " of the circuit is high enough. But this assumes a circuit of constant resistance. It is shown in section 12 (*a*) that where the circuit resistance depends on retroaction, it will not be truly a constant for all values of signal amplitude but will increase with signal amplitude when the latter exceeds a certain value. Now the centre peak in Fig. 3*b* corresponds to a maximum radio-frequency amplitude as compared with the others, and the effective circuit resistance may be so high, compared with that for the smaller amplitude, that the centre peak is, so to speak, buried under the slopes of the side peaks.

In case this explanation seems fanciful, the actual observed variation, shown in Fig. 3*c*, is given in confirmation. It refers to a retroactive coil circuit, the observations relating to which were afterwards discarded because it was discovered that the voltage amplitudes had exceeded the limit for an appreciably constant resistance. The curve is not very symmetrical, but shows the two-peaked response, with the paradoxical result that the centre or main tuning point corresponds to a minimum of modulation-frequency intensity.

The result would not occur in the absence of a resistance variation, for as the modulation frequency is decreased and the side peaks approach one another, the centre peak increases in relative intensity giving a single peaked curve as the limiting case.

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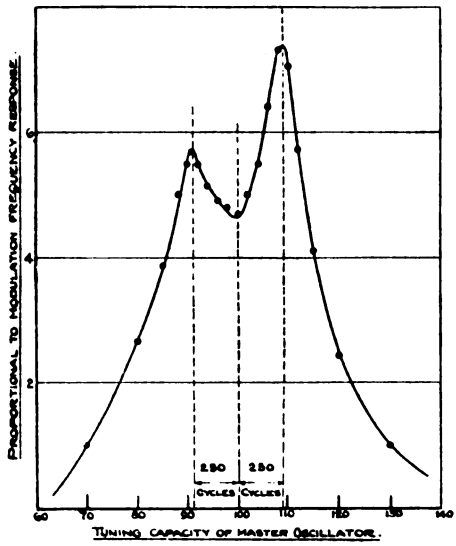

FIG. 3*c*.

7. COMPARISON BETWEEN HETERODYNE AND MODULATED WAVE RESPONSES

One of the main objects of the present investigation was to determine whether reception on circuits of very high selectivity could be expected to afford any special immunity to interference caused by the heterodyning of the carrier of the wanted transmission with extraneous excitations due to the carrier or side waves of transmissions on neighbouring frequencies. If high selectivity reception was found to possess this property, it might make it possible to increase very considerably the traffic-carrying capacity of the available radio spectrum by decreasing the frequency separation of modulated transmissions. Special attention was directed to this point, not only on account of its obvious practical importance, but because an impression had gained currency that such circuits did, in fact, exercise some sort of discrimination as between a balanced double side wave system and an unbalanced single side wave system (i.e., a heterodyne combination), giving a normal response to the one and to the other a very much smaller response than theory would indicate.

The necessary formulae for the theoretical consideration of the matter have already been established. For a modulated wave

$$e = \mathrm{E}\,(1 + \mathrm{M}\cos pt)\sin \omega t \quad \ldots\ldots .50$$

the tuning of a selective circuit to the carrier frequency gives a

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modulated radio-frequency potential difference at the rectifier of amplitude

$$V = E\left\{Y_c + M\,Y_s \cos\left(pt - \frac{\phi_1 - \phi_2}{2}\right)\right\} \quad . \quad . \quad . \quad .51$$

The response to a corresponding single side wave or heterodyne combination having the same carrier and side wave amplitude and the same frequency separation can be obtained by putting $E_1 = E$, $E_2 = (ME/2)$ and $\omega_1 - \omega_2 = p$, with corresponding substitutions for Y_1 and Y_2, in the formula 26 already developed for heterodyne reception. This gives

$$V = E\left\{Y_c^2 + \frac{M^2 Y_s^2}{4} + M\,Y_c Y_s \cos\left(pt + \phi_c - \phi_1\right)\right\}^{\frac{1}{2}} \quad . \quad .52$$

The second term in this expression will, in general, be small compared with the first, and the third will be sufficiently small to justify an approximated binomial expansion, giving as a simplified result which will be more nearly correct the higher the selectivity of the circuit,

$$V = E\left\{Y_c + \frac{M\,Y_s}{2} \cos\left(pt - \frac{\phi_1 - \phi_2}{2}\right)\right\} \quad . \quad . \quad . \quad .53$$

It is immediately apparent from a comparison of these formulae that the modulation-frequency responses for the similar double and single side waves excitation will be in the ratio of 2 : 1 in voltage, with either square law or linear rectification. This conclusion is, moreover, quite independent of the particular values of Y_c and Y_s (subject to the not very stringent conditions of the approximated calculation) and is thus independent, not only of the selectivity of the circuit, but also of its shape, provided this be symmetrical about the central frequency.

The conclusion is fairly obvious on general physical grounds, but it has been thought worth while to give an explicit formulation for the reasons already mentioned.

The result has been confirmed experimentally by a large number of measurements, both on retroactive coil circuits and on quartz crystal circuits, as described in Parts II and III.

It may be mentioned at this stage that although with the quartz circuits the ratio was found in some cases to be slightly less than 2 : 1, for reasons given in Part III, such circuits permit of an adjustable asymmetry of resonance which can be utilised for the almost total elimination of a specific heterodyne or single side wave response. The adjustment is not, however, entirely free from an adverse effect on the uniformity of overall modulation-frequency response. This is discussed more fully in Part III.

Before leaving this subject, brief mention will be made of the somewhat academic case of simultaneous excitation by a modulated wave and a single side wave of the same side wave amplitude and frequency separation. The response will clearly depend on the

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precise phase relationship between the additional single side wave and existing coincident side wave of the modulated system. If it is in anti-phase, the combination reduces to a single side wave system, and if in phase, it becomes a double side wave system with one side wave twice as large as the other. It is easily shown that, depending on these phase relationships, the response would vary between 1 and 3 times that given by the single side wave, or heterodyne, combination. The difficulty of establishing and maintaining any precise phase relationship precluded experimental verification of this result. It is, however, not of very great practical importance compared with the more general case of interference considered in the following section.

8. Excitation by Simultaneous Transmissions

(*a*) *Two modulated waves*

The interaction in reception of two (or more) modulated transmissions, the modulations being complex as in actual practice, is far too complicated a matter for precise formulation, but considerable insight into the subject can be gained by considering an ideally simplified case represented by the reception on a highly selective circuit of a pure tone modulated transmission giving an intensity

$$e = \mathrm{E}\,(1 + \mathrm{M}\cos pt)\sin \omega t \qquad \ldots\ldots\ldots .54$$

with a neighbouring transmission

$$e' = \mathrm{E}'\,(1 + \mathrm{M}'\cos p't)\sin \omega' t \qquad \ldots\ldots .55$$

it being assumed that the carrier frequency separation $(\omega - \omega')/2\pi$ is in the audible range.

Assuming that the typical receiver specified in Section 4 (*a*) is tuned to the carrier frequency of the first (wanted) transmission, the resulting amplitude of the rectifier potential difference is, as already shown,

$$\mathrm{V} = \mathrm{E}\left(\frac{1}{\mathrm{P}} + \frac{\mathrm{M}}{2\alpha}\sin pt\right) \qquad \ldots\ldots .56$$

It is necessary to determine the corresponding quantity of the second (unwanted) transmission, to which the receiver is not tuned. It could be determined exactly by the methods already indicated, but a sufficiently close approximation for present purposes can be arrived at assuming that, on account of the high selectivity of the circuit, the transformation ratio Y is effectively constant over the range $\omega' \pm p'$ at the value Y (ω'). This is equivalent to assuming that ω' lies on the " skirt " of the resonance curve of the receiver, which is substantially level over the unwanted modulator range at the height corresponding to ω'. On this basis the unwanted transmission will give rise to a rectifier potential difference of amplitude

$$\mathrm{V}' = \frac{\mathrm{E}'}{2\beta}\,(1 + \mathrm{M}'\cos p't) \qquad \ldots\ldots .57$$

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where

$$\beta = \frac{\omega - \omega'}{\omega} = \frac{b}{\omega} \quad \ldots\ldots\ldots .58$$

Thus β can be described as the "fractional carrier heterodyne frequency."

Before proceeding to the more complicated matter of simultaneous excitation, it will be well to point out the effect of the rectifier in the case of separate excitation by these transmissions. Such a case might arise in practice, for example, in the reception of a "tonic train" morse transmission where there is no sustained excitation except on the actual marking signal.

With square law rectification, the ratio of the fundamentals of the wanted and unwanted modulation frequency responses, assuming separate excitation, is

$$\frac{M' E'^2}{M E^2} \cdot \frac{\alpha P}{2 \beta^2} \quad \ldots\ldots\ldots .59$$

with linear rectification it is

$$\frac{M' E'}{M E} \cdot \frac{\alpha}{\beta} \quad \ldots\ldots\ldots .60$$

In the first case, the interference ratio is proportional to P (the power-factor of the circuit), while in the second it depends on the frequencies only, and not on the circuit power-factor. (This conclusion is, of course, only true of "high selectivity" circuits as defined in Section 6A.) Further, the interference ratio with linear rectification is higher than that for square law rectification in the ratio of $(E/E')\ (2\beta/P)$. Thus assuming that $P = 10^{-4}$, $\omega = 2\pi \times 10^5$ and $b = 2\pi \times 10^4$, so that $\beta = 10^{-1}$, the interference ratio with square law rectification, and with equal intensity amplitudes, would only be 1/2000 of that given by linear rectification. While it would be easy to overestimate the significance of such a numerical example, it seems to indicate that in some cases square law rectification might be found to have considerable advantages in respect of a silent background for morse reception.

In relation to the general problem of the simultaneous excitation of the receiver by the wanted and unwanted transmissions, special mention is made of the valuable paper by C. B. Aiken[7], in which a very comprehensive discussion is given of the rectifier interactions of two modulated transmissions. However, since Aiken takes the rectifier voltages as his starting point, it is necessary to supplement his work by means of the analysis given in the present paper in order to bring out the effect of circuit selectivity on the various interference components. Apart from thus complementing the treatment, the following discussion is based mainly on Aiken's analysis.

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The total rectifier potential difference will be obtained by combining the modulated components $V \sin \omega t$ and $V' \sin \omega' t$ (omitting certain quite unimportant phase angles), where V and V' are as already determined. The combination can be synthesised into a combined modulated amplitude given by $(V^2 + V'^2 + 2VV' \cos bt)^{\frac{1}{2}}$, where b is the angular beat frequency $\omega - \omega'$. The frequency analysis of the expression will first be considered in relation to square law rectification.

(i) *Square law rectification.*—The response is given by

$$\tfrac{1}{2}(V^2 + V'^2 + 2VV' \cos bt) \quad \ldots\ldots\ldots .61$$

The first two terms represent the simple addition of the responses due to the individual transmissions. Neglecting the relatively unimportant double modulation frequency terms, the amplitudes of the wanted and unwanted modulation frequency components are given by

$$\frac{M E^2}{2 \alpha P} \quad \ldots\ldots\ldots\ldots .62$$

and

$$\frac{M' E'^2}{4 \beta^2} \quad \ldots\ldots\ldots\ldots .63$$

In considering the practical aspect of interference, two points call for special emphasis. In the first place, the most suitable measure of the interference appears to be the ratio of the intensity of any unwanted frequency component to that of the wanted modulation. Unless otherwise stated, the term interference is used throughout in this sense of an *interference ratio.* In the second place, the significance of any given interference intensity will clearly depend on the purely psychological factor of intelligibility. The variety of practical cases makes it impossible to estimate generally the weight to be allotted to this factor, but it will probably be agreed that an intelligible form of interference will tend to be more distracting than an unintelligible interference of equal intensity.

In the present instance, the reproduced unwanted modulation frequency, generalised to cover the modulation spectrum of the unwanted transmission, will be classed as intelligible. The interference ratio is given by

$$\frac{M' E'^2}{M E^2} \cdot \frac{\alpha P}{2 \beta^2} \quad \ldots\ldots\ldots\ldots .64$$

The proportionality to P shows that this type of interference is reduced, without theoretical limit, by increase of selectivity (decrease of P). (The influence of tone correction on this conclusion is discussed in a later section.)

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The product term $2VV' \cos bt$ will give rise to a number of sum and difference tones which call for separate consideration. They are shown in the following table :—

Item.	Frequency.	Amplitude.	Ratio to wanted modulation.
1	p	$\frac{M E^2}{2 \alpha P}$	1
2	p'	$\frac{M' E'^2}{4 \beta^2}$	$\frac{M' E'^2}{M E^2} \cdot \frac{\alpha P}{2 \beta^2}$
3	b	$\frac{E E'}{2 P \beta}$	$\frac{E'}{E} \cdot \frac{1}{M} \cdot \frac{\alpha}{\beta}$
4	$p \pm b$	$\frac{M E E'}{8 \alpha \beta}$	$\frac{1}{4} \cdot \frac{E'}{E} \cdot \frac{P}{\beta}$
5	$p' \pm b$	$\frac{M E E'}{4 P \beta}$	$\frac{1}{2} \cdot \frac{E'}{E} \cdot \frac{M'}{M} \cdot \frac{\alpha}{\beta}$
6	$p \pm p' \pm b$	$\frac{M M' E E'}{16 \alpha \beta}$	$\frac{1}{8} \cdot \frac{E'}{E} \cdot M' \cdot \frac{P}{\beta}$

The third item is clearly the heterodyne beat between the two carriers. The most important point to notice about its interference ratio is that it is independent of P, i.e., of the circuit selectivity. If E' is put equal to (M/2) E and α put equal to β, the ratio becomes 1/2 as already determined for this case.

The fourth item arises from the heterodyning of the wanted side wave with the unwanted carrier. Generalised over the whole range of the wanted modulation frequencies, it produces what Aiken calls displaced (wanted) modulation spectra. It forms a part of the background noise that has been popularly described as " side-band splash." In telephony it will be unintelligible. Further, the corresponding interference ratio is proportional to P, and therefore inversely proportional to the circuit selectivity. The fifth item corresponds to similar displaced (and therefore unintelligible) modulation spectra of the unwanted modulation, due to the heterodyning of the wanted carrier and the unwanted side waves. The corresponding interference ratio is independent of circuit selectivity. Together with the sixth item, which represents unintelligible cross-heterodyning of wanted and unwanted side waves, it constitutes the remainder of the so-called " side-band splash." Of this unintelligible background noise, therefore, two important groups (items 4 and 6) are capable of reduction, to the point of practical elimination, by increase of circuit selectivity. The same applies to the reproduced unwanted intelligible modulation, but the interference components due to heterodyne interactions involving the wanted carrier, are not reduced in relative intensity by circuit selectivity.

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(ii) *Linear rectification.*—The corresponding analysis for linear rectification is much less simple, as it involves the complete frequency analysis of $(V^2 + V'^2 + 2VV' \cos bt)^{\frac{1}{2}}$. This is given in detail by Aiken in the paper already referred to, and the substitution in his formulae of the values of V and V′ given above will complete his work by including the effect of circuit selectivity. The completeness of Aiken's analysis results in somewhat complicated formulae, and for the determination of that part of the interference which is of the unwanted modulation frequency, the analysis given by E. V. Appleton and D. Boohariwalla[19] is much simpler in form and sufficiently exact for present purposes. It gives as the amplitude of the unwanted modulation frequency component

$$\frac{M'}{2} \frac{E'^2}{E} \frac{P}{4\beta^2} \quad \ldots\ldots\ldots\ldots .65$$

The amplitude of the wanted modulation frequency term is very approximately the same as that for simple reception, i.e.,

$$\frac{ME}{2\alpha} \quad \ldots\ldots\ldots\ldots .66$$

and the interference ratio is thus

$$\frac{M'E'^2}{M E^2} \cdot \frac{\alpha P}{4\beta^2} \quad \ldots\ldots\ldots\ldots .67$$

which is exactly half the corresponding ratio for square law rectification.

The form of this interference ratio is of great interest. In the first place, in spite of the linear law of rectification assumed, the interference ratio contains the *squares* of the transmission amplitudes. This emphasises the importance of the geographical factor in frequency allocations. The weaker transmission will be more subject to this interference than the stronger in the ratio of the fourth powers of the amplitudes. It is further apparent that the circuit selectivity has a two-fold effect. In the separate reception of the two transmissions the above interference ratio is, as already shown,

$$\frac{M'E'}{M E} \cdot \frac{\alpha}{\beta} \quad \ldots\ldots\ldots\ldots .68$$

This can be interpreted as representing the effect of circuit selectivity alone. It is true that the selectivity does not appear explicitly in this ratio, but it must be remembered that high selectivity, as defined in section 6A, is implicit in the assumptions on which the simple reception formulae are based. It should, however, be noted that once the circuit power-factor P has been reduced sufficiently to justify the approximate formulae, further reduction will not produce any corresponding increase in the circuit selectivity effect

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alone. In simultaneous reception the ratio is further reduced by a factor

$$\frac{1}{2} \cdot \frac{E'}{E} \cdot \frac{P}{2\beta} \quad \ldots\ldots\ldots\ldots .69$$

Writing this in the form

$$\frac{1}{2} \cdot \frac{E'}{2\beta} \Big/ \frac{E}{P} \quad \ldots\ldots\ldots\ldots .70$$

shows it to be one-half the ratio of the rectifier potential differences of the carrier waves of the unwanted and wanted transmissions respectively. (Cf. the paper by Appleton and Boohariwalla, referred to above.) This further reduction, which is explicitly proportional to the circuit selectivity, thus represents a consequence of the assumed linear rectification. The effect is well known as a fact of experience, and has been described as a " demodulation effect " of the weaker transmission by the stronger. The term is, however, somewhat misleading and it may be hoped that it will fall into disuse. It may be well at this point to correct an impression given by the earlier publications on this subject [8, 9, 10], namely, that the interference reduction by detector action only occurs when the carrier frequency difference is supersonic with respect to the detector circuit load, i.e., when the time constant of the detector circuit load is high compared with the carrier beat period. It now appears, on the contrary, that the full interference reduction effect can only be expected when the time constant of the detector circuit load (e.g., the grid leak and condenser in the case of a grid circuit rectifier) is sufficiently small for the potential changes of the carrier beat frequency to be reproduced without appreciable reduction. This fact has been pointed out by Professor E. V. Appleton in a letter published in the *Wireless Engineer* in April, 1932, and a further letter by the same writer, published in the *Wireless Engineer* in July, 1932, gives confirmatory experimental results on this aspect of the matter.

It may be of interest to illustrate these interference ratios by reference to a numerical case. We will assume equal transmission amplitudes and modulation percentages, equal modulation frequencies (1,000 c/s.), a wanted carrier frequency of 1,000 kc/s., and a frequency separation of 10 kc/s. This gives $\alpha = 0{\cdot}001$ and $\beta = 0{\cdot}01$. The circuit selectivity effect alone gives, therefore, an interference ratio of $\alpha/\beta = 1/10$. The rectifier effect is represented by a further factor of $P/4\beta$, and taking P as 10^{-4}, this amounts to 1/400. The combined effect is a ratio of 1/4000. This is not an impracticable figure.

Finally, it should be noted that the above interference ratio for unwanted to wanted modulation contains the factor P/β^2, i.e., the ratio of the circuit power factor to the square of the *fractional heterodyne beat frequency* $(\omega - \omega')/\omega$. This is, of course, the idea underlying the use of supersonic heterodyne reception with a low

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intermediate frequency as a means of obtaining very high effective selectivity with circuits of normal power-factor. The mechanism of such frequency change reception is incidentally depicted in the above analysis. With M′ put equal to zero, the second transmission can be regarded as the local heterodyne oscillator, and the beat frequency term, together with the $p \pm b$ terms, then represents the intermediate frequency modulated wave which is selected by the intermediate frequency amplifier and finally rectified at the second detector.

Returning now to the detailed consideration of the various interference components, it has been shown that the unwanted modulation interference can be reduced (in most practical cases, eliminated) by increase of circuit selectivity. This, however, is not true of the beat frequency interference ratio, the magnitude of which can be shown to be very approximately the same as for the case of square law rectification. The interference ratio for the displaced unwanted modulation spectra $(p' \pm b)$ is similarly independent of circuit selectivity. The remaining terms $(p \pm b)$ and $(p \pm p' \pm b)$ are considerably smaller than for the square law case, and are, moreover, proportional to P and can thus be reduced indefinitely by increase of circuit selectivity.

The main conclusions of the above detailed analysis of the simultaneous reception of two modulated waves can be summarised by the statement that those components of the interference which are due to heterodyne interaction not involving the carrier wave of the wanted transmission (i.e., the components of angular frequencies p', $p \pm b$, and $p \pm p' \pm b$) can be reduced relative to the wanted response without theoretical limit, by increase of circuit selectivity, whereas the relative intensities of those interference components which arise from heterodyne interactions involving the carrier wave of the wanted transmission (i.e., the components of angular frequencies b and $p' \pm b$) are independent of circuit selectivity, and cannot, therefore, be so reduced.

The part played by tone correction in this matter is discussed in Section 9.

(*b*) *Simultaneous excitation by continuous wave telegraph transmission*

The simultaneous excitation of a highly selective receiver by a number of continuous wave telegraph transmissions, together with a local oscillator for heterodyne purposes, affords a very clear illustration of the above discussion of the two main types of interference, and further points to a possible application of high selectivity reception to the separation of such transmissions. This is a matter of some practical importance, for such transmissions are now so closely spaced that in many cases (particularly at long wave-lengths) separation can only be effected by a multiplicity of tuned circuits or by the use of tuned audio-frequency amplification.

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Let it be assumed that a tone-corrected highly selective receiver is excited by a number of continuous wave telegraph transmissions, the frequency separations between successive transmissions being only a few hundred cycles per second. A suitably adjusted local oscillator of frequency f c/s. will give audible beat notes with all the transmissions lying in a band $f \pm n$, where n is the upper audio-frequency limit of the receiver. If the receiver is tuned to the local oscillator, the system is clearly analogous to that of a modulated wave with complex modulation, the only differences being that the side waves are single instead of being in pairs, and are intermittent owing to the keying. All the transmissions will therefore be fully reproduced in the receiver, with their original relative intensities. If one only is wanted, the responses due to the others must be regarded as interference, but it is interference of the carrier heterodyne type which is not reduced by the circuit selectivity. (There will, in addition, be intermittent and unintelligible heterodyne interactions between the individual transmissions, but this is interference of the kind which is reduced by the circuit selectivity.)

The transmissions can, however, be separated, for, as in the analogous modulated wave case, each will give a second and individual peak of response corresponding very approximately to resonance with the actual transmission frequency. Thus, if the circuit be tuned, not to the local oscillator, but to the wanted transmission, the wanted response will not only be greater than in the former case (with linear rectification) but the intelligible morse interference, due to the other transmissions, is now of the reducible kind (i.e., not involving any component in resonance with the circuit) and is therefore greatly reduced, if not eliminated. The various cross heterodynes amongst the unwanted transmissions are similarly reduced in intensity as in the former case and only the intermittent heterodyning between the wanted and unwanted transmissions remains ; but even this can only occur on the wanted marking signal and is not, therefore, significant as interference. Thus very high selectivity permits of the simultaneous reception of all the neighbouring transmissions if required (a " stand-by " condition) or of their separate reception, free from interference, by a slight change of tuning. This facility may be of practical value.

It may be pointed out that the decrement or power-factor limitation for morse reception, discussed by Turner and Best[21], is not necessarily an obstacle to this type of reception, for the signal frequencies can if necessary be raised by a supersonic heterodyne process. In the experimental verification of the above theory, described in Section 19 of Part III, long wave transmissions (frequencies in 15-20 kc/s.), were received and separated in this way, by frequency changing up to 80 kc/s., on a circuit with a power-factor as low as $2 - 3 \times 10^{-4}$, with scarcely perceptible " ringing."

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9. Tone-Correction

It was pointed out in the introduction that the side-band attenuation, due to reception on a highly selective circuit, can be corrected by means of audio-frequency amplification having a suitable frequency characteristic.

Before discussing this in detail, however, it will be well to clear up a point on which there seems to have been hitherto a certain misunderstanding. It is not unnatural to conclude that since a tone-corrected selective circuit gives a uniform response over the whole audio-frequency range, it is equivalent to a receiving circuit which is very flat in response over this range. This would imply that a tone-corrected selective circuit is no longer a selective circuit, and that a similar result would be obtained by "damping" the circuit with resistance.

A very simple physical argument will, however, suffice to demonstrate the fallacy in this reasoning. Suppose a selective receiving circuit to be excited by two modulated waves carrying the same modulation frequency, the carrier frequencies differing by, say, 10 kc/s. It was shown in the preceding section that if the circuit is tuned to one of the transmissions, the ratio of the response given by the modulation of the interfering transmission to that given by the wanted transmission will depend on the frequencies and on the selectivity of the circuit. If the circuit is very selective this interference ratio will be small. In fact, the ratio, or rather its inverse, can be taken as a measure of the selectivity of the circuit. Suppose now that post-rectifier tone-correction is applied. This means that every component of the audio-frequency response is multiplied by some number which is a function of the audio-frequency concerned. In the case under discussion the two responses are at the same audio-frequency and are, therefore, each multiplied by the same number. The ratio between them is, therefore, unchanged by the tone-correction. Since the selectivity can be measured by this ratio, the selectivity is unchanged by tone-correction.

This does not, of course, imply that the actual intensity of the various components of the interference are not increased by the tone-correction. They are increased, proportionally to their frequencies; but, taking as the rational basis of comparison their intensities relative to components of similar frequency in the wanted response, then there is no increase of relative intensities since the wanted components are also increased in intensity, and to the same extent, by the tone-correction.

It might, however, be argued that this equal frequency basis of comparison is not the most general or most appropriate basis, and it may, therefore, be well to examine somewhat more closely the effect of tone-correction on the interference ratios deduced in the preceding section.

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Consider, for example, that for the unwanted modulation frequency p', which apart from a numerical factor is, for both types of rectification,

$$\frac{M'E'^2}{ME^2} \cdot \frac{\alpha P}{2\beta^2} \quad \ldots\ldots\ldots\ldots .71$$

Assuming tone correction proportional to frequency, the tone-corrected ratio becomes

$$\frac{M'E'^2}{ME^2} \cdot \frac{\alpha P}{2\beta^2} \cdot \frac{p'}{p} \quad \ldots\ldots\ldots\ldots .72$$

which is greater than, equal to, or less than its original value according as p' is greater than, equal to, or less than, p. In either case, the ratio still contains the factor P and is therefore capable of theoretically indefinite reduction by increase of selectivity (i.e., decrease of P).

Taking now the most important single interference component (as far as actual intensity is concerned), i.e., the carrier heterodyne beat, the uncorrected ratio is

$$\frac{1}{M} \cdot \frac{E'}{E} \cdot \frac{\alpha}{\beta} \quad \ldots\ldots\ldots\ldots .73$$

and the corrected ratio

$$\frac{1}{M} \cdot \frac{E'}{E} \cdot \frac{\alpha}{\beta} \cdot \frac{b}{p} = \frac{1}{M} \cdot \frac{E'}{E} \quad \ldots\ldots\ldots .74$$

which is actually the value it would have on the former basis of comparison (i.e., $\alpha = \beta$) and is also the value it would have on a " flat " circuit.

Thus, whatever be the basis of comparison, the advantage of the high selectivity reception, which is the elimination of certain forms of interference, is not materially impaired by tone-correction. Further, it must be remembered that the tone-correction need not necessarily have the theoretical perfection here postulated. In practice, it may be desirable to limit it to the range of audio-frequencies required for adequate, as distinct from perfect, reproduction of the desired modulation. If, for example, the conditions of reception are such as to give rise to a 10 kc/s. carrier heterodyne beat, it may be desirable to design for tone-correction up to, say, 9 kc/s., with a steeply falling characteristic or " cut-off " beyond this point. These, however, are matters of technical detail with which the present investigation is not principally concerned.

Before leaving this part of the subject, however, it may be well to consider in a little more detail, for purposes of experimental confirmation, the nature of the tone-correction required for a simple or " circular " resonance.

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The form of the side-band attenuation for such a resonance curve is derivable from Section 6. The response is proportional to Y_s and, from (43)

$$Y_s = \frac{1}{\sqrt{P^2 + 4\alpha^2}} \quad \ldots\ldots\ldots .75$$

Since

$$\alpha = p/\omega_o$$

$$Y_s = \frac{I}{P}\frac{1}{\sqrt{1 + k^2 p^2}} \quad \ldots\ldots\ldots .76$$

where

$$k = \frac{2}{P\omega_o} = \frac{2L}{R} \quad \ldots\ldots\ldots .77$$

For such a characteristic an audio-frequency correction which can be made to satisfy the condition to almost any desired degree of accuracy is afforded by the circuit shown in Fig. 4. Subject to

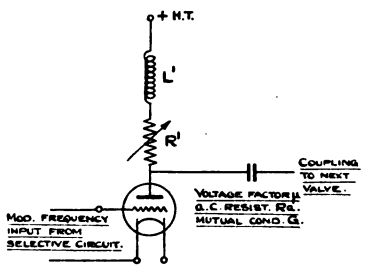

FIG. 4.

the condition that at the highest audio-frequency required, L'^2p^2 is negligible compared with R_a^2 (R_a being the slope resistance of the valve) the amplication A is given by

$$A = GR'\sqrt{1 + k'^2 p^2} \quad \ldots\ldots\ldots .78$$

where G is the mutual conductance of the valve and

$$k' = \frac{L'}{R'} \quad \ldots\ldots\ldots .79$$

The overall response of the selective circuit when followed by such an audio-frequency amplifying stage is proportional to

$$\frac{GR}{P}\frac{\sqrt{1 + k'^2p^2}}{\sqrt{1 + k^2 p^2}} \quad \ldots\ldots\ldots .80$$

and by making

$$k' = k, \text{ i.e., } \frac{L'}{R'} = \frac{2L}{R} \quad \ldots\ldots\ldots .81$$

the audio-frequency term vanishes and the response becomes uniform with respect to audio-frequency.*

* This tone-correcting circuit, of which a popular description has been published by the present writer [20] was patented by P. W. Willans in 1924 (Brit. Pat. 233,417).

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The conditions can easily be fulfilled in practice and it is thus ssible to obtain very highly selective reception of a modulated ansmission free from the characteristic distortion due to side-band tenuation.

The above is no more than a brief description of the fundamental inciples of tone-correction. It is considered from the practical d experimental point of view in Section 13 of Part II. The bject could, obviously, be developed further, but to do so would d unduly to the length of the present paper. The article[22] in e bibliography gives a theoretical account of some additional ne-correction characteristics.

. Relation of Theory of High Selectivity Reception to Radio-Frequency Filter or "Band-pass" Reception

The subject of radio-frequency "band-pass" reception as a eans of obtaining high effective selectivity, combined with equate fidelity of reproduction of modulation, has not been nsidered as such in the present survey, but it may be pointed out at, except in so far as certain of the formulæ are only applicable simple or "circular" resonance with high selectivity, the ndamental principles here enunciated are equally applicable to band-pass" reception.

A general comparison of the potentialities of the two types of ceiver is scarcely feasible, more especially as the limiting factors fecting their design are of a practical rather than a theoretical iaracter. For any actual or assumed resonance curve, however, ie methods here employed will permit of detailed estimation both response characteristics and susceptibility to interference, thus dicating which of the two main alternatives is likely to be the ore appropriate to the given conditions of reception.

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PART II

EXPERIMENTAL INVESTIGATION OF RETROACTIVE COIL CIRCUITS

II.—APPARATUS

The apparatus used in the following measurements is illustrated in the block diagram of Fig. 5. It consisted of :—

(*a*) A master-oscillator amplifier unit, with provision for modulation in the output stage.
(*b*) An audio-frequency oscillator.
(*c*) A radio-frequency rectifier valve-voltmeter.
(*d*) An audio-frequency amplifier-voltmeter.
(*e*) Tuning forks, valves, quartz crystals, etc.

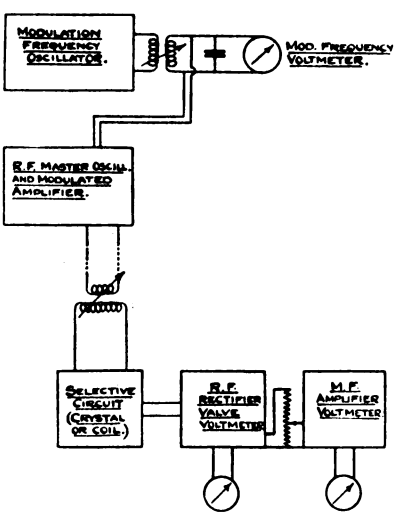
MODULATION FREQUENCY OSCILLATOR.
MOD. FREQUENCY VOLTMETER.
R.F. MASTER OSCIL. AND MODULATED AMPLIFIER.
SELECTIVE CIRCUIT (CRYSTAL OR COIL.)
R.F. RECTIFIER VALVE VOLTMETER
M.F. AMPLIFIER VOLTMETER

FIG. 5.

Brief notes on these are appended, chiefly to indicate the nature of the necessary precautions in the measurements.

(*a*) The master-oscillator amplifier unit was of normal design, but was carefully tested to ensure that it was free from appreciable frequency modulation when amplitude-modulated in the second stage. The output stage was transformer-coupled to an output coil tuned by a relatively large capacity. The variation of output

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with master-oscillator frequency was measured, in order that allowance might be made for this in calculating the modulation percentage from the modulation frequency and modulation voltage. The master-oscillator frequency calibrations were determined by a double-beat method using a fixed auxiliary oscillator, and with the same tuning forks that were used for setting the modulation frequencies.

(*b*) The audio-frequency oscillator for providing the modulation frequency voltages was a dynatron master-oscillator amplifier unit, with a loosely-coupled tuned output circuit to ensure purity of wave-form. The modulating voltage was measured by a diode valve-voltmeter, and was slightly increased with frequency to an extent which was calculated to compensate for the small side-wave attenuation effect in the output stage of the radio-frequency master-oscillator amplifier unit. The audio-frequencies were set by reference to tuning forks.

To avoid direct induction effects in the measuring apparatus it was found necessary to keep the above apparatus some ten yards or so away from the measurement circuits, the modulated output being carried along flex leads to a small coupling coil. Resonance settings, etc., were made by observing the measurement instruments with a tilted mirror and field glasses.

(*c*) The radio-frequency rectifier valve-voltmeter was arranged to operate by anode characteristic curvature, the normal anode current being balanced out of the output instrument (a pointer micro-ammeter). For square law rectification the anode voltage was 60 and the anode circuit included a tapped non-inductive resistance of 10,000 ohms. to permit of the measurement of the modulation frequency components of the anode current by the unit (*d*). This covered a range up to about one volt r.m.s. For linear rectification the anode voltage was increased to about 150 and the tapped non-inductive anode resistance to 100,000 ohms. This gave sensibly linear rectification and covered a range of voltages up to 9 volts r.m.s., in two ranges provided by means of shunts on the micro-ammeter.

Some initial difficulty was encountered in securing perfect bye-passing of the anode circuit resistances at the rather low radio-frequency involved (100 kc/s.) without using so large a capacity as to produce a falling audio-frequency characteristic. The necessary filtering effect was finally obtained by connecting a screened tuned "acceptor" circuit in parallel with the anode resistance, the tuning capacity being not greater than two or three hundred $\mu\mu$F.

The resistance tappings (ten in the 100,000 ohm case) were found necessary in order to cover the very wide range of modulation-frequency output voltages involved in the measurements.

(*d*) The audio-frequency valve amplifier-voltmeter comprised a three-valve amplifier (one resistance, one transformer coupling).

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Initially, the output valve operated a thermo-junction through a step-down transformer, but this proved too expensive in thermo-junctions, as any transient changes in the system produced momentary heavy overloads. Finally, a valve rectifier milli-ammeter was used in place of the thermo-junction.

(*e*) The audio-frequency characteristic was checked for uniformity over the range involved, and it was found on calibration that the output instrument gave a square law scale with full deflection for an amplifier input voltage of about 4 millivolts. This comparatively high sensitivity was made necessary by the fact that in some of the measurements with quartz crystals, modulation frequency currents as small as hundredths of micro-amperes had to be measured.

The overall audio-frequency characteristic from the input terminals of the radio-frequency valve-voltmeter to the output instrument of the audio-frequency voltmeter was measured. It showed a fall of some 30 per cent. or so at 4,000 cycles when the high anode circuit resistance was used, and a correction was applied for this where necessary.

It will be clear from the above notes that the measurements are not simple and contain a large number of sources of possible small error. A high degree of precision was not therefore aimed at, and is not, in fact, actually required by the nature of the investigation.

After the various sources of error had been detected and eliminated as far as possible, a very large number of measurements of response curves and resonance curves, etc., were made and showed on the whole a very satisfactory degree of mutual consistency. The results recorded are a selection of the best and most reliable of those obtained.

12. High Selectivity Obtained by Valve Retroaction

A number of measurements were made of the resonance curves and modulation frequency response curves of various coils made highly selective by means of valve retroaction, using circuit arrangements such as that shown in Fig. 6. It will be noted that the valve used for retroaction is limited to that function and is not used for rectification as well, so that the most appropriate conditions for retroaction only could be satisfied.

An examination of some of the earlier results showed a rather disconcerting divergence from the theoretical behaviour. This was eventually traced to the fact that excessive signal voltages were being used, carrying the condenser potential differences beyond the range over which retroaction remained effective. The effect was discovered by plotting the condenser potential difference against a quantity proportional to induced e.m.f., the frequency being kept constant at its resonant value. For a circuit of constant resistance

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the result should, of course, be a straight line through the origin. The result actually obtained for a given coil at a frequency of about 1,000 kc/s. is shown by the curve marked (*a*) in Fig. 7. It will be seen that there is a definite saturation effect which does not, however, set in until the output voltage (across half the coil) exceeds 3 volts r.m.s. In this case, the retroactive valve had a grid bias

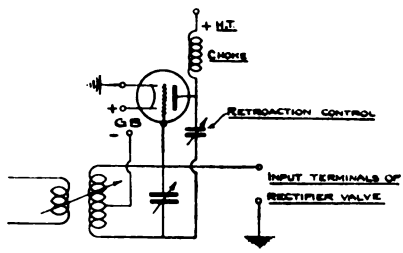

FIG. 6.

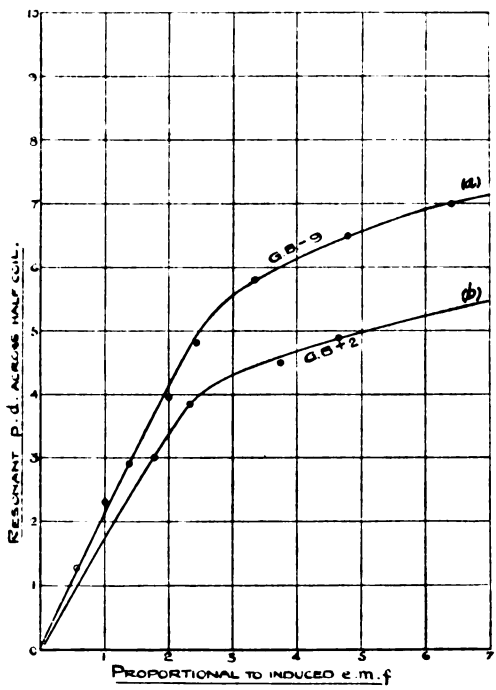

FIG. 7.

of −9 volts. In the measurements which gave the curve marked (*b*), the valve was connected as for grid rectification, with a grid leak and condenser and a grid bias of +2.

Thus, where valve retroaction is employed to obtain high selectivity, it is essential that the signal voltages shall not exceed certain limits. These are not severe limitations in practice, and

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within them it appears that the circuits do actually behave as if they had an abnormally low and sensibly constant resistance.

The following are representative of the cases examined in detail :—

	Frequency. (kc./sec.)	Tuning Cap. μμF.	Retroactive System.	Rectification.
A	1,004	1,200	Triode, PM 252, Hartley circuit, Grid bias — 9 v. H.T. 120 v. Capacity control	Linear.
B	82	300	S.G. valve used as a dynatron. Anode 60 v., S.G. 120 v. Filament control	Square law.
C	1,000	1,200	PM 252, Hartley circuit with grid leak and cond., H.T. 60 v, Grid bias + 2. Capacity control	—
D	102·5	3,000	As A but with P 240 valve, H.T. 120 v., Grid bias — 15	Linear.
E	102·2	3,000	Plain coil without retroaction	Linear.

(*a*) *Radio-frequency resonance curves*

Fig. 8 shows the radio-frequency resonance curve obtained for case A. At a height equal to $1/\sqrt{2}$ of the maximum height, the curve is only 118 cycles wide, giving for P, or $R/\omega_0 L$ ($= \delta\omega/\omega_0$ for this condition) the very low value of $1{\cdot}18 \times 10^{-4}$. This corresponds to a coil magnification factor $\omega_0 L/R$ of 8,500 to compare with 200–500 for low resistance coils without retroaction, or, say, about 100 for ordinary receiving coils. The value $1{\cdot}18 \times 10^{-4}$ for P is actually a little lower than any of the overall figures obtained for the quartz crystal circuits considered in Part III, and although the initial adjustment was rather critical, the circuit was quite stable enough for a set of measurements to be taken over a period of an hour or so.

The other cases were not adjusted to this limit of selectivity and gave values $2{\cdot}7 \times 10^{-4}$ for C (1,000 kc/sec.) and $1{\cdot}9$ and $1{\cdot}5 \times 10^{-3}$ for B and D (100 kc/sec.). These figures show that comparable values of R/L are obtained at the two frequencies. Case E refers to the plain coil used in case D, but without any retroaction. It gave a value of 222×10^{-4} for P.

The resonance curves showed a satisfactory symmetry in nearly all cases—particularly A and C (the dynatron case).

(*b*) *Modulation frequency response curves*

The circuits were excited by a series of pure modulated waves of constant intensity and constant or known modulation percentages,

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with modulation frequencies from 250–4,000 cycles, and the modulation-frequency output voltages produced in the rectifier valve anode circuit resistances were measured, and corrected where necessary for the characteristics of the apparatus. In certain of the cases the responses given when the carrier frequency was adjusted so as to bring the side waves into resonance with the tuned circuit were also measured, the circuit tuning being kept constant throughout. Fig. 9 shows a set of results for case A. The eighteen separate resonance curves shown were not actually measured in detail over

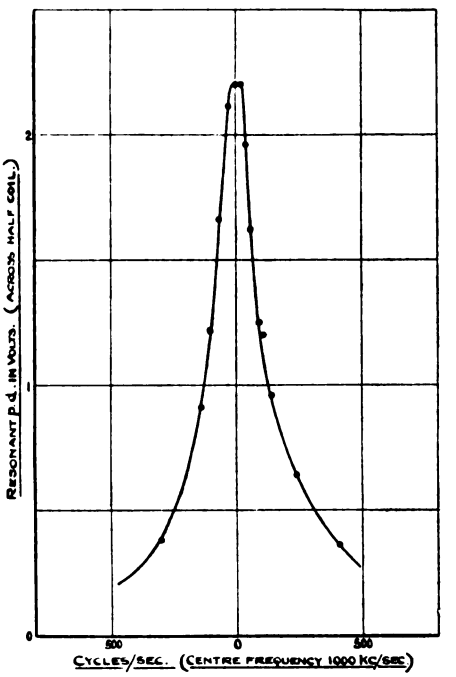

FIG. 8.

the whole course, but the maxima are actual measured values. The side peaks are higher than the centre peaks, as predicted by the theory given in Section 6 (*b*) for certain cases. This is discussed more fully in Section 12 (*c*). Theory shows that the modulation-frequency output should be inversely proportional to modulation-frequency if the selectivity is high enough. The product of output and frequency should therefore be a constant. The results shown in Fig. 10 confirm this to a reasonable degree of accuracy. The letters on the curves correspond with those of the table given above.

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(*c*) *Ratio of side wave to carrier wave peaks, and comparison of single and double side wave systems*

It was shown in 6 (*b*) that with square law rectification the ratio of the modulation-frequency output peaks corresponding to carrier

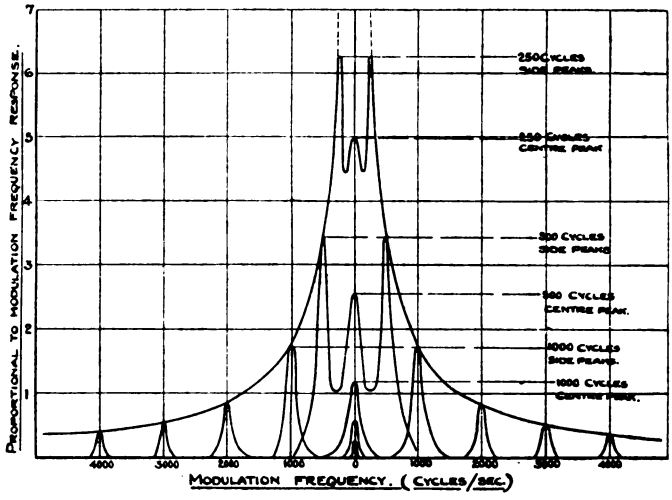

FIG. 9.

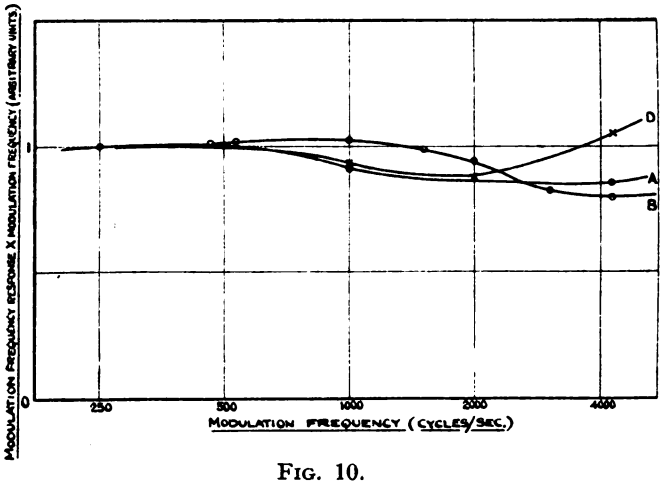

FIG. 10.

wave and side wave resonances would be as 2 to 1. This was confirmed approximately in case C, but it was later discovered that the carrier peaks had somewhat overrun the range of square law rectification and gave therefore a rather smaller output than they otherwise would have done. The ratio was found to vary between 2 : 1·3 and 2 . 1·4 for frequencies up to 4,000 cycles.

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For linear rectification the centre peaks should be lower than the side peaks in the ratio of M : 1, M being the modulation percentage. It was found impossible to satisfy the experimental conditions necessary to confirm this exactly, because the relatively small amplitude of the excitation at the side wave resonances brings the curvature at the foot of the rectification characteristic into play, and the side wave responses are therefore in general somewhat smaller than they should be, and though in every case with linear rectification the side peaks were higher than the centre peak (as in Fig. 9), the ratio of centre to side is found to be greater than M : 1. Thus in case A, the peak ratio is about 67 per cent. for all frequencies above 500, whereas the modulation percentage is 40. This result is typical of all the cases examined and was found to be consistent with the change of slope of the rectification characteristic over the regions involved in the comparison.

A more satisfactory check on the theory is given by a comparison between the modulation-frequency responses given by a modulated wave and the equivalent single side wave system, i.e., a heterodyne combination identical with that which would be obtained by eliminating one side wave from the modulated system. In this comparison, the rectification characteristic is not appreciably involved (Section 7). For the measurement it is necessary to know fairly accurately the modulation percentage or the carrier wave amplitude of the modulated system. Fortunately, the high selectivity of the circuits under examination makes this a comparatively simple matter, for it gives complete resolution of the radio frequency components of the modulated wave for all modulation frequencies above 500 cycles/sec. or so. Thus the procedure adopted was to vary the master-oscillator (carrier) frequency so as to bring each component of the modulated wave into tune with the measurement circuit, and so get a measure of its intensity, applying in subsequent calculations the necessary correction for the effect of the oscillator output tuned circuit. (For the 100 kc/s. carrier frequencies, this correction amounted to about 30 per cent. at 4,000 cycles/sec. modulation frequency). In some cases the result was checked against a calculated modulation percentage based on the measured characteristics of the master-oscillator unit and found to give a sufficiently close agreement.

The equivalent heterodyne combination was then formed by adjusting a local oscillator to give a similar side wave intensity and beat frequency.

The results in all cases confirmed the theoretical ratio of 2 : 1 for the double and single side wave systems, irrespective of the type of rectification and the selectivity of the circuits. Thus, for example, the highest selectivity circuit, with linear rectification (case A), gave values of 11·3 and 6, i.e., a ratio of 1·88 : 1 for a comparison at 1,000 cycles. As an additional check, the relative intensities of the heterodyne combination were reversed and gave

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an output proportional to 16·9, for comparison with 17·2 given by bringing one of the side waves of the modulated system into tune with the circuit. As another example, the dynatron circuit (case C) gave relative figures of 2·23 and 1·11, i.e., 2·01 : 1 (approximately square law rectification). Case F (plain coil without retroaction) was also tested in this way. In this case it was necessary to determine the modulation percentage from the oscillator characteristics owing to the very low selectivity of the circuit. This gave relative figures of 343 and 170.

13. The Response Curve of a Tone-Corrected Circuit

Finally, as a matter of practical interest, measurements were made of the response curve of a receiving circuit embodying a very highly selective retroactive tuned input coil, and an adjustable tone-correcting stage as described in Section 10 of Part I. The circuit was practically as shown in Fig. 4, but the tone-correcting impedance was actually connected as a shunt circuit to a high impedance choke in the anode circuit of the rectifier valve. This has the advantage of giving a low impedance grid circuit for the following valve, and makes possible the use of stage-separating resistances.

The input circuit was adjusted to nearly oscillation point. In this state, a commercial Vernier drive dial on the condenser did not give sufficiently fine adjustment, and a fine adjustment condenser of small capacity was connected in parallel with half the coil. The voltage limitation for the type of retroaction employed implies a small amplitude, approximately square law, rectification, but, as already pointed out, the harmonic distortion associated with this type of rectification is greatly reduced in a high selectivity circuit on account of the reduction in effective modulation percentage at the rectifier.

When tested on a broadcast transmission the circuit, as judged by ear, behaved according to plan, the response varying from excessive bass to excessive treble as the resistance of the correcting stage was decreased.

A rough measurement was made of the radio-frequency resonance curve of the retroactive circuit, and gave values varying from about 3 to 8 $\times$ 10^{-4} for P ($= R/\omega_o L$) depending on the setting of the retroaction condenser.

In measuring the response curve, the resistance was adjusted till about the same output (as measured by a valve a.c. milliammeter operated by the last valve) was obtained at 250 and 4,000 cycles. It was found that a capacity connected across the correcting inductance gave an additional factor at the higher audio-frequencies which tended to compensate for the frequency characteristic of the grid circuit rectification (which follows the same law as a tuned circuit in this respect).

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Fig. 11 shows the audio-frequency characteristic, in *db* relative to the electrical power output at 250 cycles, obtained in one of these measurements. It shows a fall of 2 *db* at 8,000 cycles, which may be compared with the fall of 10 *db* at 4,000 cycles in certain published response curves for commercial receivers.

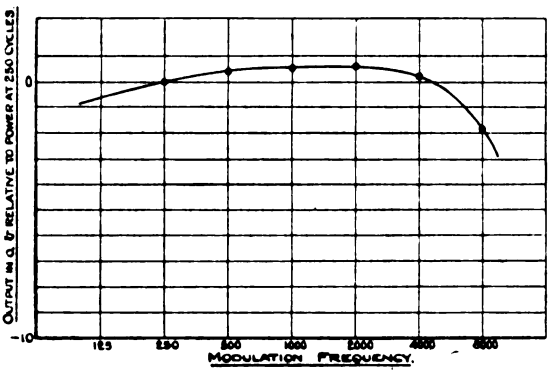

FIG. 11.

14. CONCLUSIONS TO PART II

The conclusions arrived at as a result of the experimental work described in Part II are that stable receiving circuit selectivities of the order 10^3 (i.e., circuit " power factors " $R/\omega_0 L$ of the order 10^{-4}) can easily be obtained by means of valve retroaction, using quite ordinary valve circuits, and that the behaviour of these selective circuits in relation to the reception of modulated continuous waves is in accordance with the theory given in Part I.

A similar investigation of the behaviour of receiving circuits embodying quartz crystals is described in Part III.

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PART III

THE APPLICATION OF QUARTZ CRYSTALS TO SELECTIVE RECEPTION

15. Theory

It has been shown by Butterworth[11] and Dye[12] that a quartz crystal piezo-electric resonator and its electrodes can be represented by the electric circuit shown in Fig. 12. In this figure, C_1 can apparently be regarded as determined mainly by the normal capacity of the quartz considered as a dielectric, and C_2 as a capacity due to the gap, if any, between the electrodes and the faces of the crystal.

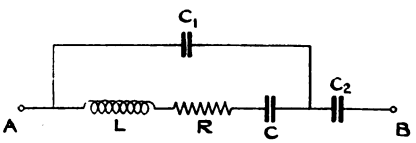

Fig. 12.

As an indication of the magnitudes of these quantities, Dye gave for a bar resonator about 7 cm. long, 0·15 cm. thick, and 0·6 cm. wide,

$$C = 0{\cdot}08\ \mu\mu F.$$
$$R = 1{,}500 \text{ ohms}$$
$$L = 160 \text{ henries}$$
$$C_1 = 8\ \mu\mu F.$$

while C_2 depends on the magnitude of the air gap and may have any value from 1 $\mu\mu$F. or so up to tens of $\mu\mu$F. For this case, the "power factor" $\omega_0 CR$ or $R/\omega_0 L$ for the resonant angular frequency ω_0 defined by $\omega_0^2 LC = 1$ is of the order $0{\cdot}3 \times 10^{-4}$, or about one-third of the lowest value obtained by valve retroaction in the experiments described in Part II. It is shown later that it may be difficult to retain this high value in an actual receiving circuit.

In what follows, the quartz resonator will be regarded purely as the equivalent electric circuit of Fig. 12, without regard to the physical properties of the quartz on which this equivalence is based.

At an angular frequency ω, the impedance of the network between the points A and B is given by

$$Z = \frac{R + j(\omega L - 1/\omega C)}{\left(1 + \frac{C_1}{C} - \omega^2 LC_1\right) + j\omega C_1 R} + \frac{1}{j\omega C_2} \quad \ldots .82$$

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The full analysis of this impedance as a function of frequency has been given by Dye[12] and will not be considered in detail here, but it is necessary to point out the principal features of its behaviour.

For simplicity it will be assumed that the air gap between the electrodes and the crystal faces is so small that the series air gap capacity C_2 can be neglected. Then Z will clearly have a minimum value (i.e., an acceptor circuit resonance) when $\omega^2 = \omega_1^2 = 1/LC$ very approximately. The denominator will, however, become very small, giving a maximum value of Z (a rejector circuit resonance), when $\omega^2 = \omega_2^2 = (1 + C_1/C)/LC_1$. With respect to an e.m.f. applied between the points A and B, therefore, the circuit behaves as a series acceptor circuit for the angular frequency ω_1 and as a rejector circuit for an adjacent angular frequency ω_2, where putting $\omega_2 - \omega_1 = \delta\omega$

$$\frac{\delta\omega}{\omega_1} = \frac{1}{2}\frac{C}{C_1} \quad \ldots\ldots\ldots\ldots .83$$

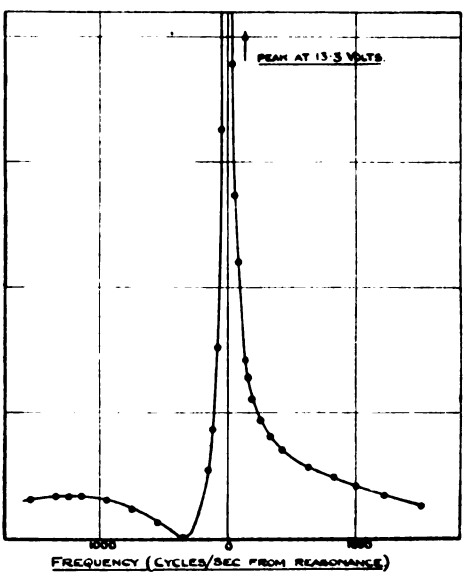

FIG. 13.

For the typical values listed above, the rejector circuit resonance will occur at a frequency about 0·5 per cent. higher than the acceptor circuit or transmission resonance.

This combination of resonances produces a curve of the shape shown in Fig. 13. Here $\delta\omega/\omega_1$ is actually 0·45 per cent. ($\delta\omega$ = about 360 cycles/sec.). The most obvious characteristic of the curve from the reception point of view is its asymmetry, which would make it unsuitable for the reception of modulated waves for the reasons given in Part I.

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In the quartz crystal receiving circuit which has been published under the name of " the Stenode Radiostat," an ingenious method has been employed for controlling the effect of the parallel capacity C_1. It is shown diagrammatically in Fig. 14.

The balancing capacity C_1' is excited by an e.m.f. in anti-phase to the e.m.f. operating on the quartz crystal, and a simple way of depicting its effect is therefore to represent it as negative capacity added to C_1, giving a resultant capacity $C_1 - C_1'$.

In the published descriptions of the Stenode circuit,[18] it is stated that the principal function of this balancing capacity is to retain the selectivity of the L R C branch of the network, unimpaired by the unselective transmission through C_1. It would seem, however, that the establishment of an approximate symmetry in the shape of the resonance curve is an even more important function as far as the reception of a modulated wave is concerned.

It has, however, an additional function which is certainly of considerable value in some special circumstances. The frequency displacement of the rejector circuit resonance, being now, on this very much simplified theory, proportional to $C/(C_1 - C_1')$, can be controlled by adjustment of C_1', and the " blind spot " in the

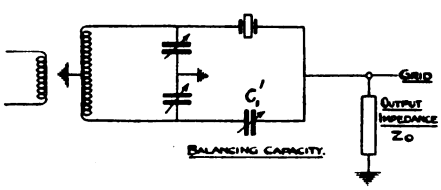

FIG. 14.

resonance curve moved to any desired frequency in the neighbourhood of the acceptor resonance, either above or below it in the spectrum. It can thus be utilised to " eliminate " a specific heterodyne interference, due to a neighbouring carrier wave, above or below the desired transmission in frequency. However, it will be apparent from the theoretical discussion given in Part I that the resulting asymmetry of the resonance curve will have undesirable consequences as far as the reproduction of modulation is concerned. Not only will a certain very small degree of harmonic distortion be introduced by the unbalancing of the modulated waves over a narrow region of frequencies centred on the interfering frequency, but the side band attenuation law will necessarily be modified in this region, and if the tone-correcting stages are appropriate to normal operation in an approximately balanced condition, then they will not give true correction when the balance is disturbed in order to eliminate a heterodyne interference. This effect will be illustrated experimentally in a following section.

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The phrase " approximate symmetry " has been used throughout in the above description. It is evident that the description has been over much simplified for purposes of qualitative explanation. In practice it is found that the precise shape of the transmission resonance curve is a very critical function of the balancing capacity C_1', and that a change of a fraction of a micro-microfarad may affect it very appreciably. Also, although by very careful adjustment an apparent symmetry of response may be obtained for some given heterodyne or modulation frequency, as judged by the equality of the side wave peak resonances, it is found impossible to secure symmetry over the whole audible range. This also will be illustrated by experimental results in a later section.

It must be realised, in this connection, that it is not the shape of the central peak that really matters in respect of modulation-frequency response, but the relative heights of the " skirts " of the resonance curve at points distant from resonance by $\pm$ modulation-frequency. The direct measurement of these " skirts " is difficult owing to the very high impedance of the crystal at frequencies so far removed from resonance, and the measurement of the audio-frequency responses at the respective side wave resonances with a modulated wave excitation is a convenient means of exploring these regions of the resonance curve.

No attempt has been made to formulate a complete theoretical description of the behaviour of the quartz crystal network as actually used in reception. A reference to Dye's paper[12] will show that the analysis of the quartz crystal equivalent circuit alone is laborious and complicated, and when the additional variables of the bridge network with the balancing condenser, and the output impedance, are also introduced, the formulation becomes unmanageably complex. Moreover, the nature of the magnitudes involved is such that the stray capacities of the system are not likely to be negligible in their effect on the behaviour of the network. These further complicate the system, and are also not easy to determine without an elaborate set of measurements.

The inferences from the experimental work described below are therefore mainly qualitative in character. It was, nevertheless, found possible to establish, in a typical case, quantitative consistency between the observed radio-frequency resonance curve and the modulation-frequency response as indicated by the general theory given in Part I, and the measurements as a whole indicate that no additional feature other than the asymmetry of the resonance curve need be introduced to explain the observed behaviour of the quartz circuits.

Nothing has yet been said of the means of associating the quartz network with the rectifier valve, i.e., the nature of the output impedance in Fig. 14. It will now be shown that this is a very important element of the system from the point of view of overall selectivity.

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16. The Effect of the Coupling Impedance between Crystal and Valve

It will be assumed, for simplicity, that the electrode capacity (C_1 of Fig. 12) is effectively neutralised by the balancing capacity C_1', and that the input tuned circuit can be regarded as a source of constant e.m.f. This will be very approximately true if its resonance curve is sufficiently flat relative to that of the crystal. The crystal network with the output impedance whereby it is coupled to the following valve is then reducible to the circuit shown in Fig. 14*a*.

This is admittedly a crude approximation to the actual circuit, but it suffices to bring out a point of considerable importance in the application of quartz crystals to selective reception.

So far as the writer is aware, the published descriptions[15, 16, 17] of the Stenode circuit show two alternatives for Z_0. In one case[16] it is shown as an ordinary parallel tuned circuit, presumably tuned to the crystal frequency. In another [17] it is shown as a high resistance (of about 1 megohm) and, since the following valve

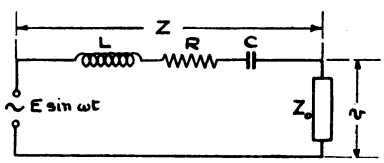

Fig. 14*a*.

is shown as a screen grid valve, it may be assumed that there is only a small input capacity (10 $\mu\mu$F. or less) in parallel with it.

A little consideration will show that neither of these arrangements is suitable unless it is desired to reduce very considerably the overall selectivity of the network as compared with that of the crystal.

For an e.m.f. **e** of angular frequency ω, the output potential difference **v** is given by

$$\frac{\mathbf{v}}{\mathbf{e}} = \frac{Z_o}{Z + Z_o} \qquad \ldots\ldots\ldots .84$$

where Z and Z_o are the impedance operators at angular frequency ω of the equivalent crystal circuit and the output impedance. Putting $Z_o = R_o + j X_o$,

$$\frac{\mathbf{v}}{\mathbf{e}} = Z_o \frac{1}{R + j(\omega L - 1/\omega C) + R_o + j X_o} \qquad \ldots .85$$

$$= Z_o \frac{1}{(R + R_o) + j(\omega L - 1/\omega C + X_o)} \qquad \ldots\ldots .86$$

The selectivity of the crystal equivalent circuit is proportional to $\omega L/R$. The overall selectivity of the crystal and output

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impedance will be very approximately proportional to $\omega L/(R + R_o)$ (since the inductive part of X_o, if any, will be very small indeed compared with ωL, L being of the order of hundreds of henries). The effect of the output impedance is thus to reduce the selectivity in the ratio of R to $(R + R_o)$, where R may be of the order 10^3 in ohms.

Now suppose Z_o to be a parallel tuned circuit. Then at the crystal frequency R_o will be very large—at least of the order 10^4—and X_o will be zero, so that the selectivity is likely to be reduced very considerably.

The measured resonance curves did, in fact, show a reduction of the order 1 : 5 when a tuned output impedance was used, as in one of the published Stenode circuits. The loss of selectivity should on this theory be much reduced if the tuned circuit has a very large tuning capacity (giving a correspondingly small value for R_o). This was confirmed by measurement.

A further disadvantage of the tuned output impedance is that the transmission ratio $\mathbf{v}/\mathbf{e}$ contains Z_o as a factor and the side band attenuation effect of the output coil will therefore also appear as a direct multiplying factor. Thus, with a tuned output circuit, we may expect a comparatively low overall selectivity and, at the same time, though this seems paradoxical at first sight, an increased *relative* side-band attenuation as compared with an untuned output impedance (i.e., an attenuation increasing more rapidly as a function of modulation frequency). This effect was also confirmed by measurement.

The use of a high resistance (such as 1 megohm) for Z_o will clearly have an even more destructive effect on selectivity, unless there is a sufficient capacity in parallel with it, but will be free from the additional side band attenuation effect.

It is necessary for the control of the mean grid bias of the valve following the crystal network, that the output impedance shall be conducting for direct current, but it is clear that it should be of low radio-frequency resistance and high radio-frequency reactance, if the crystal selectivity is to be retained without loss of sensitivity. A resistance with a parallel capacity can be used, but the most suitable coupling would seem to be a radio-frequency choke of fairly low resistance. This was not actually used in the measurements to be described, but the dependence of the selectivity on the nature of the output impedance was sufficiently confirmed by the measurements.

17. The Measurement Circuits

Two quartz crystals were examined. They will be referred to as A and B respectively. A was a bar crystal about 3·7 cm.

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long, about 3 mm. thick and about 6 mm. wide. It was loaned for the purpose by Dr. D. W. Dye, and was mounted in the adjustable air-gap holder described by him in his original paper[12]. The air gap was adjusted to as small a value as possible consistent with free movement of the crystal. The principal frequency was about 82 kilocycles per second. The other, B, was a thinner bar crystal about 3 cm. $\times$ 1 cm. $\times$ 0·2 cm., supplied by a commercial firm for use in receivers. It was mounted horizontally between plate electrodes in an evacuated glass bulb, similar to that of a screen-grid valve, the electrodes being brought out to a top terminal and to the valve pins. The principal frequency was about 102 kilocycles per second.

The measurement circuit was as shown in Fig. 15. In the diagram the output impedance is shown as a tuned circuit. This was actually a small receiving coil of about 5 mH inductance, tuned to the necessary frequencies by a variable condenser with a maximum capacity of about 1,000 $\mu\mu$F. It was mounted in a metal screening case and was set in a direction of minimum coupling

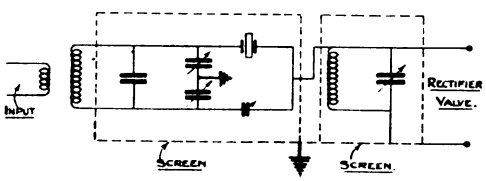

FIG. 15.

with the input coil. It was confirmed that the direct induction in this coil was negligible. In certain of the measurements a coil of much lower inductance (about 1 mH) was used in this place, tuned by a larger capacity (some 3,000 $\mu\mu$F. or so). In others, a grid leak resistance with a shunting capacity of 100 to 500 $\mu\mu$F. was used.

Measurements were made of the radio-frequency resonance curves; the modulation-frequency response, both for carrier and side wave resonance, as a function of modulation frequency, for a constant modulation percentage; the comparison between the audio-frequency responses given by a modulated wave and the equivalent single side wave (or heterodyne) system. In the above measurements the balancing condenser C_1' was adjusted to give a "balanced" response with respect to some particular modulation-frequency—usually 250, 500, or 1,000 cycles—i.e., it was adjusted so that equal audio-frequency responses were given with a heterodyne oscillator set at $\pm n$ with respect to a constant carrier frequency, which was set at the crystal resonance, or, alternatively, so that equal responses were obtained at the side wave resonances of a

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suitable modulated wave. Other measurements were made of the radio-frequency resonance curves and audio-frequency response curves with the balancing condenser set so as to eliminate some specific heterodyne interference, with a view to confirming the mechanism of this elimination and its probable effect on quality.

In all cases the circuit selectivity was high enough to permit of the direct measurement of the amplitudes of the carrier and side waves concerned by bringing these separately into resonance with the crystal circuit.

A selection of the results will now be given, in relation to the specific features investigated.

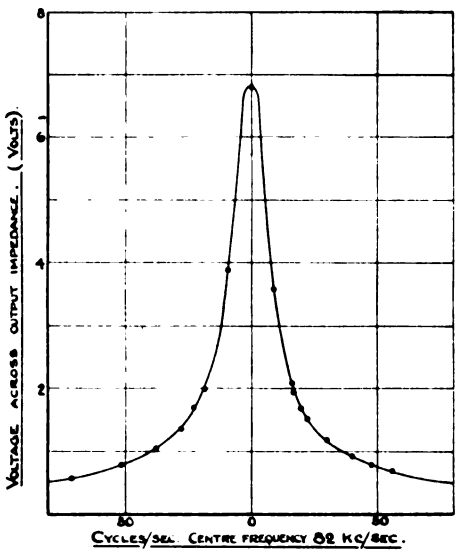

FIG. 16*a*.

18. EXPERIMENTAL RESULTS

(*a*) *Resonance curve of balanced circuit*

Fig. 16*a* shows the radio-frequency resonance curve for crystal A, coupled to the rectifier valve by means of a resistance of 2·6 megohms, shunted by a capacity of about 100 $\mu\mu$F. This gives, at the frequency 81·7 kilocycles per second, resistance and reactance components of about 160 ohms. and 20,000. ohms. respectively for Z_0 (Fig. 14*a*, Section 16). It appears that 160 ohms. is fairly small compared with the resistance of the crystal circuit, as an increase of the shunt capacity across the coupling resistance to 600 $\mu\mu$F. did not appreciably sharpen the resonance curve. The balancing

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capacity was adjusted by reference to a heterodyne frequency of $\pm$ 250 cycles per second.

A comparison of this resonance curve with that in Fig. 13 shows the effect of the balancing capacity C_1' in eliminating the "rejector circuit" dip in the curve and so obtaining an approximate symmetry, at least in the neighbourhood of the central resonance. The approximate value of P (the effective "power factor" of the circuit), judged by the width of the curve at $1/\sqrt{2}$ of the maximum height, is $1{\cdot}7 \times 10^{-4}$, which is rather high compared with the figures given by Dye. The cases are not, however, very closely comparable.

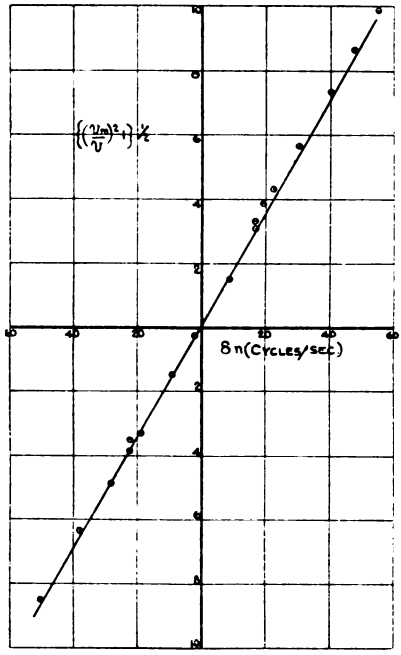

FIG. 16*b*.

Fig. 16*b* shows the corresponding "rectified" resonance curve $\pm \{ (Y_m/Y)^2 - 1\}^{\frac{1}{2}} = \partial\, \omega\, (2/P\, \omega_o)$. Its rectilinearity shows that the circuit has the characteristics of simple resonance, as far as the centre peak is concerned, and the more accurate value of P deduced from the slope is $1{\cdot}4 \times 10^{-4}$.

(*b*) *Effect of coupling impedance on selectivity*

A large number of radio-frequency resonance curves were taken with various forms of output or coupling impedance Z_o (Section 16).

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Some typical results are given in the following table. The letter indicates which crystal is referred to :—

No.	Frequency. kilo-cycles per second.	Output impedance.	$P \times 10^4$ (from res. curve).
A1	81·7	2·6 meg., 100 μμ F. in parallel	1·7
A2	81·7	2·6 meg., 600 μμ F. in parallel	1·7
A3	81·7	5 mH coil, tuning capacity about 600 μμ F.	7·7
A4	81·7	,, ,, ,,	6·5
B1	102	2·6 meg., 100 μμ F. in parallel	3·2
B2	102	1 mH coil tuned by capacity about 3,000 μμ F.	4·6
B3	102	5 mH coil, tuning capacity about 400 μμ F.	20·9
B4	102	,, ,, ,, Balance condenser disconnected	31

The values of P, determined by direct measurement of the width of the resonance curves, cannot be considered as very accurate, but they serve to confirm the general conclusions with regard to the effect of Z_0 on selectivity. In every case the use of a tuned circuit of high " dynamic " resistance results in a considerable lowering of the selectivity. A coil with large tuning capacity, however (cf. B2) causes a smaller loss of selectivity, but will, of course, be associated with a correspondingly smaller overall sensitivity.

(*c*) *Calculation of modulation-frequency response, from the data of the measured radio-frequency resonance curve*

The bulk of the measurements made on the behaviour of the crystals were concerned with certain relative values only. In view, however, of certain published statements to the effect that the measured modulation-frequency response of a quartz crystal unit differs from the value which would be calculated from its measured characteristics[13], it was thought worth while to make an absolute measurement of the response and compare it with the observed radio-frequency curve and the measured characteristics of the rectifier. Agreement can only be expected on the basis of simple resonance curve theory over the range in which the resonance curve remains of this character, i.e., symmetrical, and of the shape which gives a straight line when " rectified " in the sense of Section 4A. In the case listed as A1 above, it was found that this condition would be satisfied for modulation frequencies up to 500 cycles per second at least, and two measurements were therefore made, at modulation frequencies of 250 and 500 cycles per second. The carrier and side

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wave amplitudes of the incident field were determined by bringing these respectively into resonance with the circuit and noting the actual voltage readings given by the rectifier-voltmeter. The actual side wave rectifier voltage amplitudes were then calculated in accordance with the theory given in Part I, using the measured value of P from the radio-frequency resonance curve. Calling the side wave rectifier voltage amplitude v_s, and representing the rectifier characteristic by $i_c = f(v)$, where v is the amplitude of the applied radio-frequency voltage, then the modulation-frequency component of the rectifier output current is given in R.M.S. value by $\sqrt{2}\, v_s\, (\partial i/\partial v_c)$, where v_c is the carrier wave amplitude, and the R.M.S. value of the alternating voltage in the rectifier anode circuit resistance R ohms is simply R times this current. The actual response was determined by a calibration of the audio-frequency amplifier-voltmeter under the conditions of operation in the measurement. The results were as follows :—

	Modulation frequency (cycles/sec.).		
	250	500	
Calculated response ..	$2{\cdot}6 \times 10^{-3}$	$5{\cdot}0 \times 10^{-3}$	volts,
Measured response ..	$3{\cdot}2 \times 10^{-3}$	$4{\cdot}0 \times 10^{-3}$	R = 1,100 ohms.

Considering the complexity of the measurement and the smallness of the quantities involved, the agreement is satisfactory, and there would seem to be no discrepancy between the observed and calculated behaviour when all the relevant factors are taken into account.

(*d*) *Effect of balancing capacity on radio-frequency resonance curves*

Figs. 17*a* and 17*b* show the effect on the shape of the resonance curves of an adjustment of the balancing condenser for the purpose of eliminating a heterodyne interference at $\pm$ 500 cycles/sec. respectively. This is, of course, an extreme case and one not likely to arise in practice, but is included to illustrate the description of the effect of the balancing condenser given in Section 15. The illustration of the effect by reference to a more practical case, e.g., the elimination of a 10 kilocycles per second heterodyne note, is hardly practicable owing to the difficulty of getting a measurable radio-frequency response at frequencies so remote from resonance.

The effect of this adjustment on the audio-frequency response is discussed in Section 18 (*e*).

(*e*) *The variation of modulation-frequency response with modulation frequency*

It was shown in Part I that with a receiving circuit of very high selectivity having a symmetrical resonance curve, the modulation-frequency response to a modulated wave would be inversely proportional to modulation frequency, and that the curve of response $\times$ modulation frequency should, therefore, be a straight line parallel

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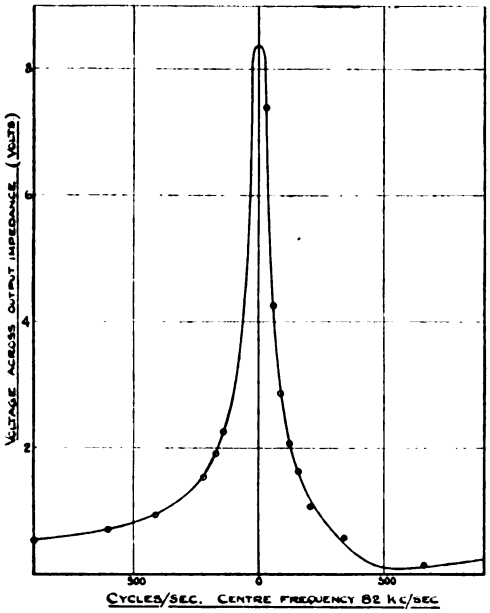

FIG. 17*a*.

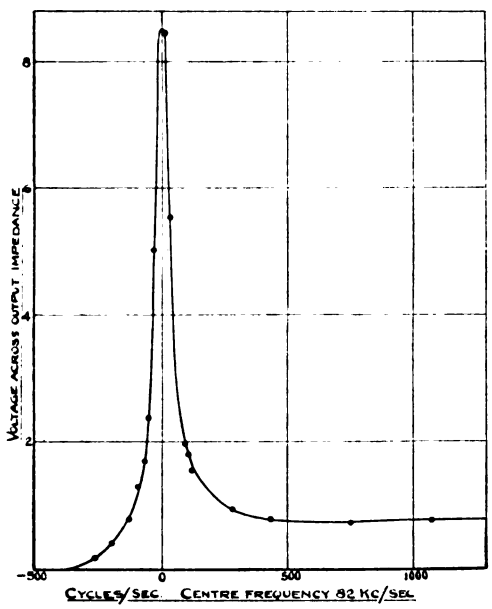

FIG. 17*b*.

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to the frequency axis. This was confirmed in Part II for a retroactive coil circuit.

It was explained in Section 15 that one of the objects of the balancing capacity C_1' is to obtain a symmetrical resonance curve for the crystal unit. It was also stated that the shape of the resonance curve is a critical function of C_1' and that in practice it is very difficult, if not impossible, to obtain complete symmetry of the curve over the whole range of audible frequencies. If a particular heterodyne frequency, say 500 cycles/sec., is selected and the capacity adjusted to give a symmetrical response for $\pm$ 500, it will be found that the " rejector dip " in the resonance curve is

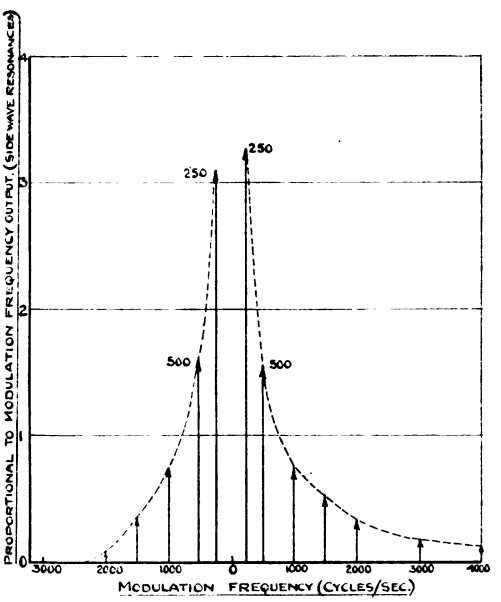

FIG. 18.

still located in the audible band outside 500 cycles and that a pronounced asymmetry occurs in that region. The curve in Fig. 18, for example, shows the modulation-frequency responses for crystal B, with a capacitative output circuit, corresponding to the side wave resonances (i.e., the side waves in each case brought into tune with the crystal). Balance was first obtained at 1,000 cycles, and it will be seen that the response is symmetrical up to this limit. There is, however, so marked a dip at $-2{,}000$ onwards that the responses on that side became too small to measure. It might appear that balancing for a higher heterodyne frequency, e.g., 4,000 cycles, would meet the case, but it was found that the dip might then be found inside this limit at some lower frequency. In fact, no really practicable method was found for securing

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symmetry over the whole range of audible frequencies, though it seemed by comparison of a large number of curves that a 1,000 cycle note was, on the whole, the most reliable medium for balancing.

To an unsymmetrical curve, the simple inverse law cannot be expected to apply. Curves (1), (2) and (3) of Fig. 19 show that it does not, in fact, apply. In these cases the output impedance was capacitative, giving high selectivity (P of the order 10^{-4}) and

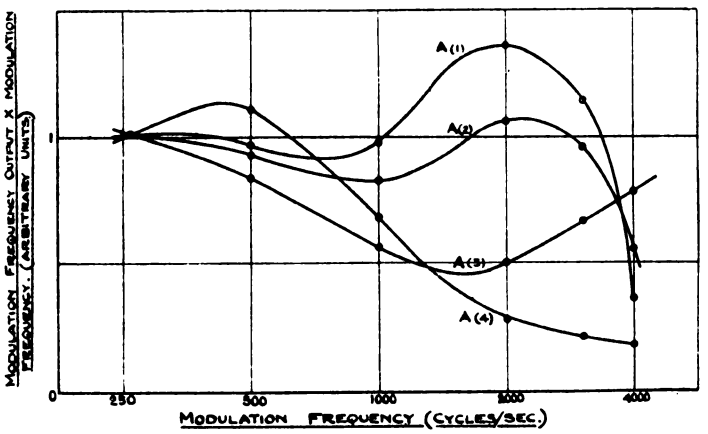

FIG. 19.

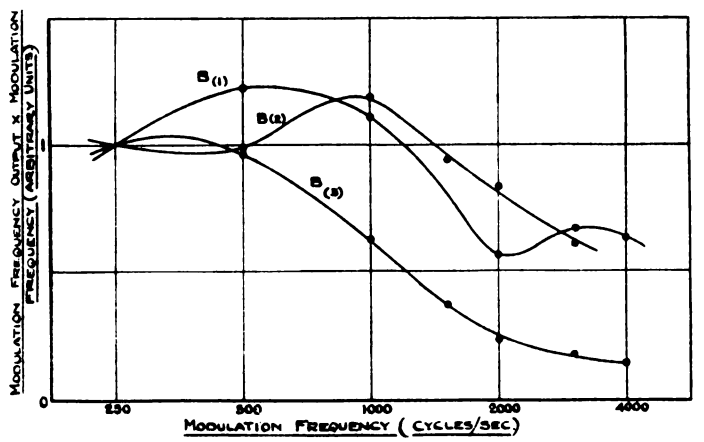

FIG. 20.

balance was adjusted at 250 cycles. In case 4 the same crystal was used, balanced at 500 cycles, but with a tuned circuit output impedance, giving a much lower selectivity. It shows, on the whole, a steeper side band attenuation, i.e., steeper than $1/n$, in spite of its lower selectivity, as predicted by the theory given in Section 15.

The curves of Fig. 20 were obtained under similar conditions for crystal B. It happens that these were all balanced at 1,000

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cycles, and curves (1) and (2) hold up better than the corresponding curves of Fig. 19. Curve (3) was taken with a tuned coil output circuit, and shows the increased relative side-band attenuation in spite of lower selectivity, as in the corresponding case for crystal A.

Curves (1) and (2) of Fig. 21 are related to the curves of Figs. 17*a* and 17*b* respectively, and show the effect on the modulation-frequency response of an adjustment of the balancing condenser to eliminate heterodyne notes at −500 and +500 cycles/sec. respectively. The actual response at 500 cycles modulation-frequency was comparable with that for the balanced case, showing that the elimination of one side-band was compensated by an increase in the other due to the lower overall selectivity resulting

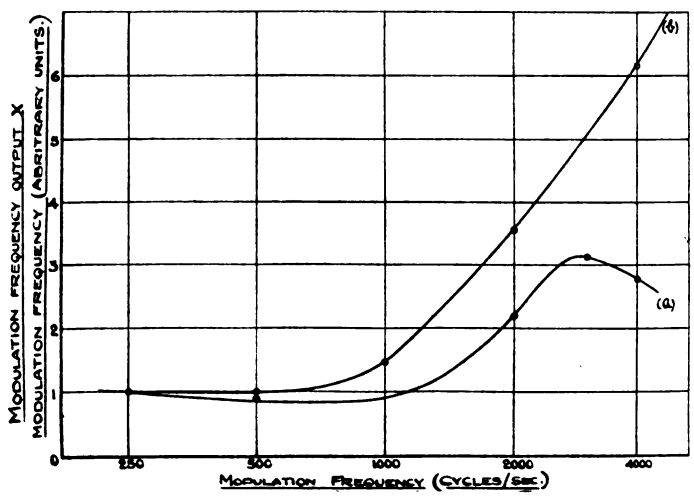

FIG. 21.

from the unbalancing of the system. This point was further tested by a measurement of the responses to a ± 1,000 cycle heterodyne note with the corresponding ∓ 1,000 cycle heterodyne eliminated by the balancing condenser. They were proportional to 29·8 and 12·5. With the system balanced for this frequency the responses were each equal to 9, showing that the elimination of a particular heterodyne interference on one side tends to produce an increased response to similar frequencies on the other side of resonance.

The elimination effect is clearly spread over an appreciable width of the spectrum, but, as already explained, the 250 cycle elimination is not a case which would arise in practice and is included merely as illustrative of the general theory. The curve

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of Fig. 22 is a more practical case. Here the balancing condenser for crystal B was adjusted to eliminate a 4,000 cycle heterodyne note. It was not, in this case, found possible to trace the radio-frequency resonance curve over the range of frequencies involved owing to a power limitation in the source. Here, again, the dip is spread over an appreciable part of the modulation spectrum. It is possible, however, that as the elimination frequency is increased, the dip in the response curve will tend to become narrower, but the apparatus used, though of a fairly high order of overall sensitivity, was not sensitive enough to make exact measurements at these higher modulation frequencies, where the output becomes exceed-

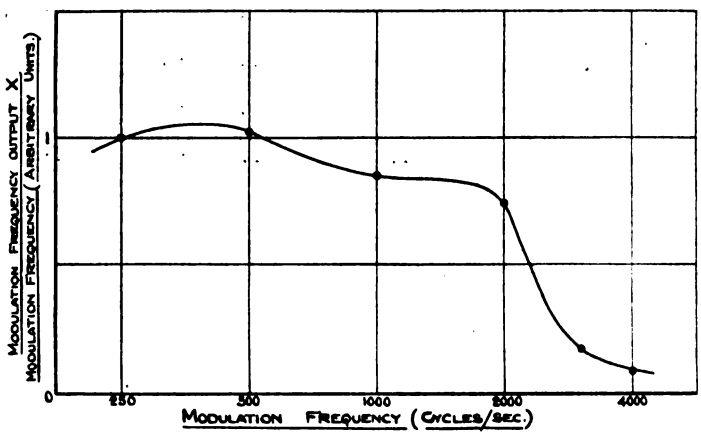

FIG. 22

ingly small—less than 10^{-8} ampere in alternating current, which is the lower limit of sensitivity of the apparatus.

(f) Comparison of single and double side-band response

The comparison of single and double side wave response was made by the same methods as those employed in the retroactive coil experiments already described, i.e., the direct measurement of the radio-frequency components of the modulated wave concerned by bringing each into resonance with the crystal, and the formation of the equivalent single side wave or heterodyne combination by means of a local oscillator adjusted to give an induced voltage equal to that of the original side waves.

In every case it was found that the ratio of the responses was approximately 2 : 1 for the modulated wave and equivalent single side wave systems, in accordance with the accepted theory, though in some cases the ratio was found to be rather less than 2 : 1. The

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following set of results, as an example, refer to crystal A with a capacitative output circuit (i.e., P of the order 10^{-4}) :—

Excitation.	Rectifier radio-frequency voltage	Audio-frequency output voltage. (arbitrary units).	
1,000 cycle mod. wave, carrier set at crystal frequency	7·2	2·3	
1,000 cycle mod. wave, lower side wave at crystal frequency	1·5	2·4	
1,000 cycle mod. wave, higher side wave at crystal frequency	1·5	4·6	
Heterodyne, carrier at crystal frequency, local oscillator at — 1,000 cycles	7·2	0·86	Mean 1·08
Heterodyne, carrier at crystal frequency, local oscillator at + 1,000 cycles	7·2	1·3	
Heterodyne, local oscillator at crystal frequency, carrier at + 1,000 cycles	1·5	2·3	
Heterodyne, local oscillator at crystal frequency, carrier at — 1,000 cycles	1·5	4·5	

The mean value for the heterodyne responses, i.e., 1·08, gives the ratio of 2·1 : 1 for the double to single side wave systems. The agreement of the second and third figures with the sixth and seventh, respectively, serves to confirm the equivalence of the modulated wave and heterodyne systems, for the remote side waves in 2 and 3 will contribute a negligible component to the audio-frequency response.

The above result is not quite representative for, in most cases, the double side band response was a little lower than would be expected. Thus, for example, an earlier measurement with crystal A gave a ratio 1·7 : 1. In particular, measurements with crystal B gave a lower than theoretical value, varying from 1·5 : 1 to 1·8 : 1, but the effect described in the following section lays these results open to a certain doubt.

In addition to the above measurements, in which the single side wave was a pure continuous wave adjusted to the appropriate amplitude and frequency, some additional determinations were made in accordance with an experimental scheme originally proposed by Professor E. V. Appleton. The carrier frequency of a modulated wave was first adjusted to resonance with the quartz crystal circuit, and the modulation-frequency response was measured. The carrier frequency was then displaced from resonance to an extent equal to twice the modulation frequency, and another carrier wave of intensity equal to that of the original carrier was adjusted to resonance with the quartz crystal circuit. The total audio-frequency response under these conditions contains a modulation-frequency component due to the interaction of the new carrier wave with the nearer side wave of the modulated wave, a double modulation-frequency component due to the interaction of the two

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carrier waves, a component of three times the modulation frequency due to the interaction of the new carrier and the further side wave of the modulated wave, and a small additional component of modulation frequency derived from the displaced modulated wave only. The last mentioned component will be negligibly small. The only significant component of modulation frequency is thus derived from the single side band system formed by the new carrier wave and the nearer side wave of the displaced modulated wave. This single side wave modulation frequency response was measured for comparison with that given by the original double modulated excitation. The response measurements were made on an audio-frequency voltmeter system sharply tuned to the modulation frequency, so that all other frequency components gave a negligible response.

In each case the modulated wave was displaced first to one side and then to the other side of resonance, in order that due account might be taken of any asymmetry in the quartz crystal circuit resonance curve. The following is a representative selection of the results obtained. Crystal B was used, with a resonant frequency at 82 kc/s. In the table of results, excitation D.S.W. means double side wave (i.e., modulated wave at crystal resonance), S.S.W. 1 or 2 means single side wave, due to displacement of modulated wave to twice the modulation frequency below or above resonance, with a local oscillator set at resonance.

Modulation frequency: Cycles per Second.	Excitation.	Modulation frequency response. (Voltage in arbitrary units).	Ratio of double to single side wave.
3,000	D.S.W.	10·5	—
	S.S.W. 1	10·2	1·03 } Mean
	S.S.W. 2	3·46	3·06 } 2·05
3,000	D.S.W.	8·0	—
	S.S.W. 1	3·0	2·67 } Mean
	S.S.W. 2	5·0	1·60 } 2·13
2,000	D.S.W.	10	—
	S.S.W. 1	5·3	1·90 } Mean
	S.S.W. 2	5·2	2·10 } 2·0
1,000	D.S.W.	9·80	—
	S.S.W. 1	4·35	2·25 } Mean
	S.S.W. 2	5·10	1·92 } 2·1

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These results are seen to be quite consistent with the similar determinations by the former method, and serve the additional purpose of demonstrating that it is immaterial whether the interfering oscillation be a pure sine wave from a single frequency source, or a side wave component of a modulated wave.

The results, as a whole, show quite clearly that the use of a quartz crystal for obtaining abnormally high selectivity, though it may have other advantages, cannot be expected to confer any special immunity to heterodyne interference. There was no theoretical reason for anticipating that it would do so, but it has been thought necessary to establish the fact on a basis of measurement, for reasons given in Section 7 of Part I.

(*g*) *An apparent amplitude effect*

Crystal B seemed to behave quite regularly in circuits embodying a tuned circuit output impedance. In some cases, however, when a somewhat lower capacitative output impedance was used, implying correspondingly high voltages in the crystal for the same output voltages, it was found that if the side wave was brought into resonance with the crystal, then the carrier, and then the further side wave, the third reading of the radio-frequency voltage was very much smaller than the first (less than a half) instead of being equal to it as it should have been. If the crystal was left quiescent for a period of 15 minutes or so, a reversal of the above cycle gave the same result in the opposite sense. A series of observations taken over a period of 20 minutes or so confirmed the conclusion indicated by the first set, that the state of the crystal was affected by large voltage amplitudes of the order required to give output voltages up to six or seven volts, and that its behaviour was not quite reproducible and depended on the length of time of application of the carrier voltage. For some reason which is not quite clear, the response to the larger (carrier) voltage seemed reasonably constant, but the subsequent response to the smaller (side wave) excitation was profoundly affected.

The fact that behaviour became normal again after a lapse of time seems to indicate some thermal effect, but the point was not investigated, as crystal A did not behave in this way, and it appears therefore to be a matter of detail concerning crystal dimensions and type of mounting.

It would not necessarily be of any significance in the practical application of the crystal to reception, as the general behaviour was not profoundly modified. Also the voltages concerned were probably somewhat larger than those which would occur in practice. However, it made the crystal unsuitable for measurement purposes except at small amplitude, and brings out the possibility of overloading a crystal and thus affecting its transmission characteristics in a quantitative sense.

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19. Continuous Wave Telegraphy Reception with a Quartz Crystal Receiving Circuit

With a view to obtaining a qualitative confirmation of the conclusions given in Section B of Part I, more particularly in relation to telegraphy reception, the quartz crystal circuit used in the foregoing measurements was connected to a rectifier and an audio-frequency amplifier with a suitable tone-correcting stage. It was preceded by a simple rectifier stage having in the anode circuit a radio-frequency transformer with secondary tuned to the crystal frequency. The input circuit of this first rectifier was tuned to the signal frequency required, and a local oscillator was adjusted so as to transform the signal frequency to that of the crystal, as in ordinary supersonic heterodyne reception. The assembly was first tested by the reception of one of the B.B.C. transmissions (London Regional) and found to behave, qualitatively, in accordance with theory. The input circuit was then tuned to about 20 kc/s., and a second local oscillator (which will be referred to as the heterodyne oscillator in order to distinguish it from the above mentioned frequency change oscillator) was adjusted so as to give audible beat notes with Rugby, Nauen, and a number of other telegraphy transmissions in the neighbourhood of 20 kc/s. Adjustment of the frequency change oscillator to about 100 kc/s. gave the necessary signal frequency change to about 80 kc/s. (the crystal frequency). This is, of course, the ordinary supersonic heterodyne process, but is somewhat unusual in that the difference frequency is higher than the signal frequency. (It may be mentioned that reception was equally well obtained with the frequency change oscillator at about 60 kc/s., the sum frequency, $60 + 20$, being produced in this case.) By variation of the frequency change oscillator it was clearly possible to bring either the local heterodyne oscillator or individual signal transmissions into resonance with the crystal circuit.

The observed behaviour was found to be quite in accordance with the theoretical discussion given in Section 8 of Part I. With the local heterodyne oscillator brought into resonance, all the transmissions were received simultaneously (generally four or five), together with a certain amount of background noise. By a very small change of the local frequency change oscillator (necessitating a fine adjustment condenser of small capacity) the individual transmissions could be brought successively into resonance, and it was found possible to receive them separately, practically free from interference and, incidentally, unusually free from background noise.

The case is, of course, somewhat analogous to the reception of a continuous wave modulated by a number of separate and distinct tones. As pointed out in Section 6 (*b*), each tone will be received at three different settings of a highly selective receiver, corresponding

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to resonance with the side waves and the carrier. One setting (the carrier resonance) will be common to all the tones; but each of the other two will be peculiar to the particular tone concerned. Thus, at the central setting all the tones will be heard simultaneously. At the side settings the tones will be separated.

This type of reception may have useful potentialities in practice. It affords a means of selecting from among closely packed telegraphic transmissions, even on low radio frequencies, and also permits, by a small change of the tuning adjustment, of the simultaneous reception of all transmissions included in a range of frequencies equal to the audio-frequency range of the receiver.*

As a point of some theoretical interest it may be mentioned that, in the simultaneous reception setting, the morse reproduction was quite sharp and "clean cut," in spite of the abnormally low decrement of the crystal circuit. With the setting for separate reception, on the other hand, there was a faint but quite noticeable trace of "woolliness," as is expected in morse reception in very low decrement circuits. Even at high hand-speed, however, the reception was perfectly clear and well defined. This is, of course, an incidental advantage of the stepping up of the signal frequency from about 20 to about 80 kc/s. Turner and Best[21] give about 3×10^{-3} as the lowest permissible receiving circuit power-factor for reception at 15 to 20 kc/s. In the present instance the power-factor was probably not more than a tenth of this figure.

* It has since been pointed out to the writer that the conditions of the above experiment are not representative of an extreme case where a transmission is required to be received in the presence of an interfering transmission of much greater intensity. In such cases it is possible that high single circuit selectivity would not give adequate separation.

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SUMMARY AND CONCLUSIONS

Part I gives a comprehensive analytical formulation of high selectivity reception within the limits imposed by the assumption of quasi-stationary conditions. It is realised that the wide field of transient conditions, more particularly in relation to impulsive modulation of a continuous wave, remains to be explored.

The experimental work on retroactive valve circuits described in Part II serves the double purpose of confirming the theory and establishing the reliability of the methods of measurement employed.

The experiments on quartz crystal circuits described in Part III were found to present considerable practical difficulties, due mainly to the very great range of magnitudes involved, and the large number of significant variables.

The experimental work, as a whole, does not, obviously, cover all the cases of interference likely to arise in practice. To do so would involve a costly and complicated elaboration of equipment if really reliable results were to be obtained. The objective has been the more economical one of checking a general theory by experimental confirmation of certain critical points. The fundamental assumptions being thus established, the more powerful and elastic methods of analysis can be applied to the investigation of special cases of practical importance.

The main conclusions that have been arrived at on the relation between high selectivity and interference can be summarised as follows :—

In relation to a desired transmission assumed to consist of a carrier wave with associated side waves due to telephonic, telegraphic, or other modulation, the total interference can be classified into two groups. The first includes all interference which would exist (though not in general with the same intensity) even if the carrier wave of the wanted transmission were completely suppressed. The most obvious and familiar example is the reproduction in the receiver of the direct modulation of a neighbouring and interfering transmission, e.g., the reproduction of the Mühlacker programme in a set tuned to the London Regional transmission. This group also contains the intermittent heterodyne noises due to the interaction of the side waves of the desired transmission with the carrier wave or side waves of any other transmission. (These form part of the sibilant background noise, popularly described as "side-band splash," which is sometimes encountered in the reception of neighbouring broadcast transmissions.)

The second group comprises all interference which would disappear completely if the wanted carrier were removed. It

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includes, that is to say, all interference which is due to heterodyne interaction between the carrier wave of the wanted transmission and waves from all sources other than that of the wanted transmission, this interaction arising from the operation of the detector or other non-linear conducting elements in the receiver.

The most familiar example of this type is the steady-heterodyne beat note due to the carrier of a neighbouring transmission, but it also includes the intermittent noises due to the heterodyning of the desired carrier with the side waves of neighbouring transmissions. (These intermittent components form the remainder of the " side-band splash.") The whole of the second type of interference may be conveniently referred to as carrier heterodyne interference, since it arises from a variety of heterodyne pairs, each containing the desired carrier frequency as one element.

It may be pointed out that all reproduction as effective output response, audible or otherwise, of radio-frequency electric waves depends upon what is essentially a heterodyne process, and the above distinction into two groups therefore resolves itself into a distinction between heterodyne interference involving the desired carrier wave, on the one hand, and all other forms of heterodyne interference, on the other hand.

The reason for dwelling at such length on this classification of interference is that, as shown both theoretically and experimentally in the present survey, there is a fundamental practical distinction between them in relation to a selective receiver. Interference of the first type can be reduced to almost any desired extent within certain practical limits of circuit and transmission frequency stability, by increasing the selectivity or " sharpness of tuning " of the receiver. This process involves a high degree of distortion of the desired modulation on account of the well known " side-band cutting " effect, but this distortion can be eliminated by means of post-rectification audio-frequency amplification having a suitably designed tone-correcting characteristic. Assuming that interference is properly measured as the ratio of undesired to desired response, it can be shown that this process of post-rectifier tone-correction does not affect the selectivity of the receiver in the above sense.

The efficacy of high circuit selectivity in eliminating interference of this type is due to two important factors. The immediate consequence of the high selectivity is a correspondingly high disparity between the desired and undesired frequency components of the total voltage operating on the terminals of the rectifier. This disparity itself further enhances the overall selectivity of the reception, for the nature of the process of rectification, whether " square-law " or " linear," as applied to complex wave forms such as those which arise from the simultaneous reception of a number of transmissions, is such as to increase the predominance of the stronger components in the output. In relation to " linear "

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rectification, this feature has been described as a " demodulation effect " on the interfering transmissions, but the term is somewhat misleading, for the effect so described is no more than a normal consequence of the assumed type of conductivity on an electromotive-force made up of two or more modulated waves.

In regard to interference of the second type, i.e., carrier heterodyne interference, the position is totally different. It is shown theoretically that such interference, measured as already indicated in terms of the ratio of undesired to desired response, cannot be reduced by circuit selectivity, and is, in fact, independent of circuit selectivity except in certain special cases where a controllable asymmetry of the resonance curve of the receiver is used to produce a virtual blindness over a certain narrow band of the radio spectrum on one side or the other of the tuned frequency. The limitations of this special device are discussed in the survey.

The investigation of this second type of interference resolves itself ultimately into the comparison of the response of the receiver to comparable double and single side wave excitation with the same carrier frequency. It is shown in the present survey that this ratio can be expected to be 2 : 1 in voltage, irrespective of circuit selectivity. The theory is confirmed very closely by measurements on retroactive coil circuits.

Certain early experiences with receivers in which an abnormally high selectivity was obtained by means of the forced piezo-electric oscillation of quartz crystals seemed to indicate that the behaviour of such circuits was not in accordance with accepted theory, and that they exercised a marked discrimination against the single side wave excitation, giving, in consequence, a much higher ratio than 2 : 1. If this were so, then the use of such circuits would clearly afford a means of reducing the interference of both types and might permit of a corresponding increase in the traffic-carrying capacity of the wireless medium.

The point has been tested by a large number of quantitative measurements. Briefly stated, the results show that the behaviour of such quartz crystal circuits of very high selectivity is quite in accordance with accepted theory if due allowance be made for a certain degree of asymmetry in the resonance curve. This asymmetry, being controllable, permits of the elimination of certain specific interferences over a narrow frequency band, but, in general, it results in a carrier heterodyne interference ratio which is actually slightly higher than that given by a normal selective circuit having a symmetrical resonance curve. In other words, the double to single side band ratio is usually found to be slightly less than 2 : 1 and not considerably greater as had originally appeared.

Thus the experimental work shows that both retroactive coil circuits and quartz crystal circuits behave in accordance with theory in relation to heterodyne interference, and that neither can

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be expected to give any reduction of carrier-heterodyne interference ratios as a consequence of their high selectivity. This statement applies to all forms of receiving circuit characterised by the symmetry of resonance which is desirable for the reproduction of double side wave systems.

In addition to the above general conclusions relating to interference, certain other matters of detail are discussed in the survey.

The great relative enhancement of the carrier voltage in high selectivity reception results in very low modulation percentage at the rectifier terminals. This has the desirable effect of reducing the harmonic distortion due to curvature, if any, in the " dynamic " or rectification characteristic of the rectifier. Tone-correction will increase the relative intensities of any harmonics so produced, but there remains a considerable nett advantage. There is, in any case, a great gain in the fact that high selectivity reception permits of rectification at a lower amplitude level, thus reducing the radio-frequency amplification required together with any cross-modulation effects due to non-rectilinearity in the radio-frequency stages.

For the practical realisation of highly selective reception of the kind here considered there are, in the first place, the familiar alternatives of direct reception at the carrier frequency or supersonic heterodyne reception. Both cases offer further alternatives in the means of obtaining the requisite sharpness of the over-all radio-frequency resonance curve, the principal distinction being between cases in which the selectivity is due mainly to a single circuit of abnormally low decrement, obtained either by the use of valve retroaction or by means of a quartz crystal, and cases in which the over-all sharpness of resonance is due to the combined action of a number of circuits of normal decrement with sufficiently loose coupling, either direct or through the agency of valves, to give a single-peaked symmetrical resonance curve. It should be noted that where multi-circuit selectivity is employed the appropriate tone-correcting characteristic will tend to be somewhat more complicated in form than that required for a single circuit of abnormally low decrement.

The actual choice between the above alternatives will depend upon the conditions of reception involved and upon a number of similar practical considerations which lie, for the most part, outside the scope of the present survey, but it has been pointed out to the writer by members of the research staffs of the Royal Aircraft Establishment, Farnborough, and of H.M. Signal School, Portsmouth, that the over-all resonance curve of a suitable combination of circuits of normal decrement may be superior to that of a single circuit of abnormally low decrement in relation to strong interference on frequencies remote from that of the wanted transmission.

In the case of the reception of broadcast transmissions the supersonic heterodyne method has the advantage that by using a low intermediate frequency a correspondingly high value of the

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fractional modulation frequency is obtainable, making it possible to obtain a high degree of effective selectivity with circuits of normal decrement.

In any case where the supersonic heterodyne method is employed, a quartz crystal can be used at the intermediate frequency as a single circuit of abnormally low decrement.

As already stated, the quartz crystal has an inherently asymmetrical resonance curve, and even with a balanced bridge connection it appears impossible to avoid a residual asymmetry of response. One consequence of this asymmetry is the production of a small degree of harmonic distortion, irrespective of any rectification process, but the extent of this distortion is likely to be negligible in practice. A more important consequence is that the law of the reduction of modulation-frequency response with modulation frequency depends appreciably on the nature and degree of the asymmetry, and in general the response falls off more rapidly than as the inverse first power of the frequency. The controllable asymmetry of the resonance curve affords a means of eliminating a specific heterodyne interference which falls within the audible range, but the shape of the response curve is very considerably modified by this adjustment and if properly corrected for the balanced condition will no longer be corrected for the unbalanced condition. It is, of course, arguable that the resulting non-uniformity in response may be preferable to the disturbance caused by the heterodyne note. This is a matter of individual opinion.

As against the above it must be stated that a higher degree of stable and constant single circuit selectivity can be obtained by a quartz crystal circuit than by any simple and practicable valve retroaction circuit at present in use. It must be remembered, however, that there has not hitherto been any great concentration of attention on the problem of attaining constant and abnormally high selectivity by means of valve retroaction, and the results described in this paper, obtained by quite ordinary means, are sufficient to indicate that it is unlikely that the limit has yet been reached in this direction. If it is eventually found possible to secure selectivity comparable with that of quartz in valve-controlled circuits operating at the carrier frequency, such circuits will have the advantage of variable frequency of operation, so that a supersonic heterodyne construction, though it may be desirable on other grounds, will no longer be necessitated by a fixed filter frequency. It should also be possible to ensure symmetry of resonance, as there need be no inherent source of asymmetry.

Assuming a suitable choice of the means of producing the requisite sharpness of the over-all radio-frequency resonance curve, the investigation indicates that the combination of this high circuit selectivity with tone-correction appropriate to the transmissions concerned, possesses properties which make it a valuable addition to the available alternatives in the design of receiving circuits.

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It is, however, considered necessary to emphasise the fact that while the theory of the subject shows that a high degree of freedom from interference can be obtained without sacrifice of quality, nevertheless, in practice, the permissible degree of selectivity for the maintenance of any given standard of quality will be strictly limited by the attainable stability of operation of the receiving circuit and of the transmissions concerned. (It should be noted that the above limitations may be expected to become more severe as the frequency is increased. In the present investigations, no measurements have been made at frequencies higher than about 1,000 kc/s.) Furthermore, the detailed analysis given in this survey makes it clear that a satisfactory performance of apparatus based on these principles will require correctness of design and precision in manipulation.

The work was carried out at the National Physical Laboratory, and the author wishes to acknowledge the encouragement and helpful comment of Dr. R. L. Smith-Rose throughout the course of it. He has been assisted by Mr. B. J. Bryne of the Wireless Division in the experimental work and analysis of observations. The author is very much indebted to the late Dr. D. W. Dye and to Mr. P. Vigoureux for the loan of quartz crystals and for guidance and assistance in certain parts of the work on crystals; to Dr. S. Brydon of the Signals Experimental Establishment, Woolwich, for much interesting information on the same subject; and to Mr. F. S. Barton, of the Royal Aircraft Establishment, Farnborough, and Mr. L. S. Alder of H.M. Signal School, Portsmouth, for valuable comment and a suggestion relating to multi-circuit arrangements.

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DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH

RADIO RESEARCH

SPECIAL REPORT No. 13

VALVE OSCILLATORS OF STABLE FREQUENCY

A CRITICAL SURVEY OF PRESENT KNOWLEDGE

BY

F. M. COLEBROOK, B.Sc., D.I.C., A.C.G.I.

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PREFATORY NOTE

THE communication-traffic carrying-capacity of the radio medium is limited in part by the practically attainable :ability of the frequency of the carrier-waves emitted by wireless elegraphy stations of all kinds.

As far as the larger fixed stations are concerned, a very high egree of frequency-stability is attained by the well-known methods f quartz-crystal or tuning-fork control, or by frequency-control by ieans of small-power master-oscillator circuits of special design.

Such methods are not, however, of universal applicability. n certain types of small-power mobile stations, for example, uch as installations on ships or on aircraft, considerations of ost, space and special conditions of operation, such as provision or changes of frequency, may necessitate the use of very simple elf-maintained oscillating circuits as the direct source of emission part, therefore, from the theoretical importance of the subject, t is desirable from a practical point of view that there shall be s complete as possible an understanding of the frequency-ehaviour of simple valve-maintained oscillators, and of the ieans of securing the highest possible degree of frequency-tability in such oscillators.

An investigation of this subject is at present being carried ut under the direction of a Committee of the Radio Research Board constituted as follows :—

Mr. F. E. Nancarrow (*Chairman*).
Mr. E. B. Moullin.
Mr. G. Shearing.

As a desirable preliminary to the investigation, a thorough urvey of the literature of the subject has been carried out for he Committee by Mr. F. M. Colebrook, and the result of this urvey has been embodied in a critical *résumé*.

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The *résumé* has been drawn up in two parts. The first part consists of an essay on the subject as a whole, summarising the most important general information at present available and illustrating it by reference to typical forms of oscillating circuit.

The second part consists of abstracts of the more important published work on the subject including such patent specifications as have been described in technical periodicals. The abstracts are representative rather than exhaustive. That is to say, an endeavour has been made to include typically important papers dealing with each aspect of the matter, rather than to cover all the relevant papers that have appeared. The papers are abstracted singly or in natural groups. Comments are appended to each abstract. These comments are not to be regarded as criticism of the work described, but are intended rather to bring out as far as possible the relation of each contribution to the subject as a whole.

Finally, it is desired to emphasise that the subject concerned is still in a state of fairly rapid development. Neither in the essay nor in the annotated abstracts is there any suggestion of finality. In fact, one of the principal objects of the *résumé* was to try to find out, in the light of previous work, the most promising directions in which to pursue the matter further. In the hope that it may serve this purpose equally for other workers in the same field, it has been decided to publish the *résumé* in its present form.

A. G. LEE,
(*Chairman Radio Research Board*).

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VALVE OSCILLATORS OF STABLE FREQUENCY

A CRITICAL SURVEY OF PRESENT KNOWLEDGE

PART I

1. CLASSIFICATION OF CAUSES OF FREQUENCY-VARIATION IN OSCILLATORS

IN attempting any classification of the causes of frequency-instability in valve-maintained oscillators, the difficulty is immediately encountered that an oscillator system with its various components and its environment, constitute essentially a single system with a considerable degree of interaction between its elements. A distinction can, however, be drawn between frequency variations due to changes of a purely electrical character and those due to changes of the physical configuration of the system, even though these are not necessarily independent. This classification will be adopted in the present discussion and variations due to purely electrical changes of the system will first be considered.

2. THE DEPENDENCE OF OSCILLATION-FREQUENCY ON THE CONDITIONS OF MAINTENANCE

2.1. The Maximum or Stationary Value of the Oscillation-Frequency

It is well known that the natural frequency of a free damped oscillation in an electric circuit consisting of inductance, capacitance and resistance (L, C, and R) in series is given by

$$\omega = \left(\frac{1}{LC} - \frac{R^2}{4L^2}\right)^{\frac{1}{2}} = \left(\omega_0^2 - \frac{R^2}{4L^2}\right)^{\frac{1}{2}} \quad \ldots\ldots(1)$$

Let it be assumed that means are available whereby R can be gradually reduced to zero or even made negative. As R is reduced ω increases continuously toward the value ω_0, which it reaches when $R = 0$. Under the same conditions the oscillations are undamped and are therefore sustained. If R be further reduced (*i.e.*, made negative), the oscillation will have negative damping (*i.e.*, will increase in amplitude) and the frequency falls

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in value. Thus ω_0, or $1/\sqrt{LC}$, is a maximum or stationary value of the angular frequency of oscillation of a circuit of inductance L and capacitance C, with respect to variation of resistance.

2.2. Analytical Representation and Physical Interpretation of the Maintenance of the Stationary Value of Frequency

Oscillation can be maintained in an L–C–R circuit by inducing in it an e.m.f. proportional to current and suitably phased. Assuming for simplicity a linear relationship, *i.e.*,

$$e = \{a(\omega) + jb(\omega)\}\, i,$$

sinusoidal oscillation of angular frequency ω is represented in vector form by

$$\left(R + j\omega L + \frac{1}{j\omega C}\right) i = e = \{a(\omega) + jb(\omega)\}\, i \quad \ldots (2)$$

or

$$\left[\{R - a(\omega)\} + j\left\{\omega L - \frac{1}{\omega C} - b(\omega)\right\}\right] i = 0 \quad \ldots (3)$$

This implies

$$R = a(\omega) \quad \ldots \text{(maintenance condition)} \quad \ldots (4)$$

$$\omega L - \frac{1}{\omega C} = b(\omega) \quad \ldots \text{(frequency condition)} \quad \ldots (5)$$

A condition of oscillation with stable amplitude implies that the valve conditions which produce the effective negative resistance term $-a(\omega)$ must of necessity adjust themselves to satisfy the maintenance condition $R - a(\omega) = 0$, at least as a mean value equilibrium, whatever the value of ω.

In general $b(\omega)$ will be a function of the constants of the valve and of the auxiliary components of the valve network, and also of the quantities L, C and R. The oscillation-frequency will therefore depend, generally to an appreciable extent, on the valve constants and on R unless the circuit can be so arranged that

$$b(\omega) = 0 \quad \ldots (6)$$

in which case the oscillation frequency will be that corresponding to the stationary value $\omega_0 = 1/\sqrt{LC}$. Circuits can, in fact, be so arranged as to satisfy this condition for an assumed linear state of operation, and those described in abstracts 1, 2.1, 2.2, 2.3 and 2.4 are of this kind. McLachlan has pointed out, however, (14) that the simple linear relationships postulated in this argument are immediately violated by the actual state of oscillation, and implies that this divergence between the actual and postulated conditions deprives this design principle of any practical value. His main contention must be admitted. On

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the other hand, it is not unreasonable to suppose that the application of this design principle should lead at least to some improvement in frequency-stability with respect to maintenance-conditions, and the proceeding is in fact quite definitely justified by practical results, even though in some cases there is a considerable quantitative discrepancy between the actual and calculated values of the stabilising elements of the circuit.

Considering the stabilised condition in physical terms, the relation $b(\omega) = 0$ implies that the driving self-induced e.m.f. must be in phase with the current. There is here an analogy with mechanical simple harmonic oscillatory systems maintained in oscillation by a sinusoidal driving force proportional to velocity and in phase with it.

(It should be noted that an impulse of infinitely short duration, acting at the instant of maximum velocity, is equally valid as a means of imparting energy and thus sustaining an oscillation without affecting frequency. *Cf.* Section 3.5 and abstract 21.)

As already stated, many of the special circuit arrangements described in the abstracts are, in effect, practical means of realising the above conditions for inherent frequency stability. Their nature can best be illustrated by reference to a simplified theory of two well-known types of valve maintained circuit, *i.e.*, the dynatron circuit and the back-coupled tuned anode-circuit.

2.3. Simplified Theory of Valve-Maintained Oscillation, and Means of Frequency-Stabilisation

2.3.1. *Dynatron type*

This term is generalised to cover any system of the type illustrated in Fig. 1, where the driving impedance Z_a has a negative resistance component.

Writing Z for the "rejector" impedance of the oscillatory circuit, the maintenance-condition takes the very simple form

$$Z_a + Z = 0 \quad \ldots\ldots\ldots\ldots (7)$$

since for this condition no e.m.f. is required to maintain the current i_a. For convenience this can be written

$$\frac{1}{Z_a} + \frac{1}{Z} = 0 \quad \ldots\ldots\ldots\ldots (8)$$

In the simple case when Z_a is a pure negative resistance of magnitude R_a, equation (7) becomes

$$Z = R_a \quad \ldots\ldots\ldots\ldots (9)$$

which shows at once that the frequency of oscillation is that for which the rejector circuit impedance Z becomes a pure resistance, *i.e.*,

$$\frac{1}{R_a} = \frac{1}{Z} = \frac{1}{R + j\,\omega L} + j\,\omega c = \frac{R}{R^2 + \omega^2 L^2} + j\left(\omega C - \frac{\omega L}{R^2 + \omega^2 L^2}\right) \cdot (10)$$

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For comparison with the standard form of equation (3), equation (10) can also be developed in the form

$$\left(R + j\omega L + \frac{1}{j\omega C}\right) - \left(\frac{R_a}{1 + \omega^2 C^2 R_a^2} + \frac{1}{j\omega C}\,\frac{1}{1 + \omega^2 C^2 R_a^2}\right) = 0 \quad (10A)$$

or

$$\left(R - \frac{R_a}{1 + \omega^2 C^2 R_a^2}\right) + j\left(\omega L - \frac{1}{\omega C} + \frac{1}{\omega C}\,\frac{1}{1 + \omega^2 C^2 R_a^2}\right) = 0 \quad .(10B)$$

From 10 or 10B can be obtained the well-known dynatron equations:—

$$R_a = \frac{L}{CR} \quad \ldots\ldots .(\text{maintenance condition}) \ldots\ldots .(11)$$

$$\omega^2 = \frac{1}{LC}\left(1 - \frac{CR^2}{L}\right) \quad .(\text{frequency condition}) \quad \ldots\ldots .(12)$$

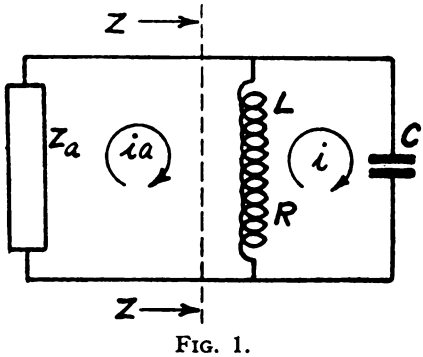

FIG. 1.

It will be seen from equation (12) that the frequency is theoretically independent of the valve constants. (In practice it will not be so for reasons to be considered more fully below.) The frequency has not, however, the ideal stationary value ω_0. To obtain this stationary value it is necessary that Z_a shall have a reactance component equal and opposite to that of Z at the frequency ω_0. In order to satisfy the maintenance condition for frequency ω_0 we must have

$$-\frac{1}{Z_a} = \frac{1}{Z(\omega_0)} = \frac{1}{R + j\omega_0 L} + j\omega_0 C \quad \ldots\ldots .(13)$$

$$= \frac{-R}{j\omega_0 L}\,\frac{1}{R + j\omega_0 L} \quad \ldots\ldots\ldots .(14)$$

or

$$Z_a = -R_a + jX_a = -\frac{\omega_0^2 L^2}{R} + j\omega_0 L \quad \ldots\ldots .(15)$$

i.e.,

$$R_a = \frac{\omega_0^2 L^2}{R} = \frac{L}{CR} \quad \ldots .(\text{maintenance condition}) \ldots\ldots .(16)$$

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and

$$X_a = \omega_0 L \quad \ldots\ldots \text{(stability condition)} \quad \ldots\ldots (17)$$

Thus the stable frequency ω_0 is theoretically obtained by inserting an inductance equal to L in the connection to the negative resistance. (This conclusion is modified somewhat, though not essentially, by any electrode-capacitance present.)

It is also interesting to note that if a resistance R is included in the capacitance branch, the rejector impedance at frequency ω_0 is

$$\frac{1}{\bar{Z}} = \frac{1}{R + j\omega_0 L} + \frac{1}{R + \frac{L}{j\omega_0 C}} = \frac{1}{R + j\omega_0 L} + \frac{1}{R - j\omega_0 L} \quad (18)$$

$$= \frac{2R}{R^2 + \omega_0^2 L^2} \quad \ldots\ldots\ldots (19)$$

This is a pure resistance, and the driving of this circuit by a pure negative resistance should therefore give the stationary frequency ω_0. This result appears to have been first published by MÖLLER in 1919 (abstract 5.2). It is also mentioned by GROSZKOWSKI (abstract 13).

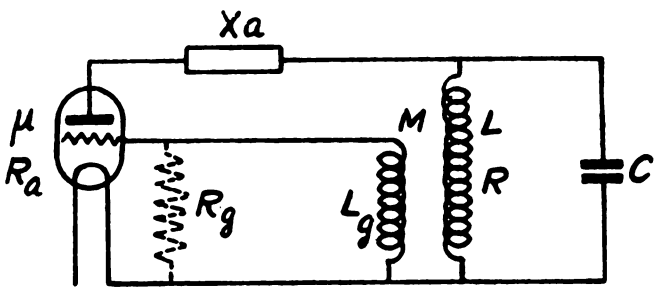

FIG. 2.

For reasons considered more fully below, the actual behaviour of the dynatron type of circuit is less simple than this approximated linear analysis would suggest.

Abstract 2.4 describes an experimental investigation of a dynatron or negative resistance circuit on the lines indicated above. The observed behaviour is qualitatively as predicted by the theory, but there is a considerable discrepancy between the observed and calculated values of the stabilising components. This is probably attributable to a departure from the simple "linear" mode of oscillation postulated in the theory, and also to a variation of interelectrode capacitance with conditions of operation.

2.3.2. *Back-coupled tuned anode-circuit type*

The simplified analysis of the circuit shown in Fig. 2 is quite familiar, but it can be put in a new form which facilitates the

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physical interpretation of the operation and illustrates some of the experimental results described in the abstracts.

It is assumed that the effect of grid-current can be represented by a resistance R_g across L_g. For generality, a reactance X_a is included in the anode connection. Putting

$$Z_g = R_g + j\omega L_g \text{ and } Z_a = R_a + jX_a$$

the complete condition of maintenance can be developed in the form

$$\underset{(a)}{\left\{R + j\left(\omega L - \frac{1}{\omega C}\right)\right\}} + \underset{(b)}{\left\{\frac{M^2\omega^2}{Z_g}\right\}} + \underset{(c)}{\left\{\frac{j}{\omega C\,(1 + j\omega C Z_a)}\right\}} - \underset{(d)}{\left\{\frac{j\mu M\omega R_g}{Z_g\,(1 + j\omega C Z_a)}\right\}} = 0 \quad \ldots (20)$$

which states that the resultant or total effective series impedance of the tuned circuit, including the effect of coupling to the valve, shall be zero, and therefore requires no e.m.f. to maintain oscillation.

The four groups of symbols represent respectively :—

(*a*) the series impedance of the tuned circuit alone ;

(*b*) an increase of resistance and decrease of positive reactance due to grid-current ;

(*c*) an increase of resistance and change of reactance due to the shunt load of the valve (and X_a) on the tuned circuit ;

(*d*) a " driving impedance " consisting of a negative resistance term and a reactance term (depending on μ, R_g, R_a and X_a).

Thus connection to the valve introduces two main increases of resistance (due to R_g and R_a) which are balanced by a negative resistance term depending on μ. It also produces appreciable changes in the reactances of the tuned circuit. The desirable condition is that the reactance components of groups (*b*), (*c*) and (*d*) shall be individually or jointly as small as possible, and the resistance component equal to — R as a total.

Abstracts Nos. 1 to 4 and 19 are essentially concerned with means whereby the reactance terms of groups (*a*), (*b*) and (*c*), or analogous terms appropriate to other types of circuit, and the effect of these terms on frequency, may be minimised or eliminated. The various means available may be briefly noted.

(i) *Elimination of grid-current.*—The minimisation of grid-current, either by means of suitable bias or, more generally, by means of automatic bias provided by a grid-leak arrangement, is obviously desirable, in itself or in association with other means of stabilisation. Abstract No. 6 is mainly devoted to this aspect of the matter.

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As $R_g \to \infty$, the second group of factors, (*b*), tends to zero, and, since $Z_g \to R_g \to \infty$, the effect of grid-conductivity is also eliminated from the fourth group, and the equation takes the simpler form

$$\left\{R + j\left(\omega L - \frac{1}{\omega C}\right)\right\} + \left\{\frac{j}{\omega C\,(1 + j\omega C Z_a)}\right\} - \left\{\frac{jM\omega\mu}{(1 + j\omega C Z_a)}\right\} = 0 \quad \ldots\ldots\ldots (21)$$

(ii) *High value of anode-circuit resistance.*—The advantage of using a valve of high internal resistance, or, alternatively, of including a high resistance in the anode connection, can be illustrated most simply by reference to equation (21), in which grid-current is assumed to have been eliminated. It will be assumed that X_a is O. Development of the second and third groups of equation (21) gives as the resistance and reactance components

$$\frac{1}{\omega C}\frac{1}{1 + \omega^2 C^2 R_a^2} - j\frac{R_a}{1 + \omega^2 C^2 R_a^2} - \frac{M\omega^2\mu C R_a}{1 + \omega^2 C^2 R_a^2} - j\frac{M\omega\mu}{1 + \omega^2 C^2 R_a^2} \quad \ldots\ldots\ldots (22)$$

In general, $\omega^2 C^2 R_a^2$ will be appreciably greater than unity, especially when R_a is large or is increased by an external resistance. Also we can put $\omega^2 C = 1/L$ as an approximation for present purposes, so that (22) becomes

$$\omega L\frac{L}{C R_a^2} - \frac{L}{C R_a} - \frac{M\omega^2\mu L}{R_a} - j\frac{M\omega\mu L}{C R_a^2} \quad \ldots\ldots\ldots (23)$$

This makes apparent the advantage of increasing R*a*. There is, of course, a limit to the process, since the negative resistance term also decreases with R_a, but it should be noted that it only diminishes as μ/R_a, whereas the corresponding reactance term diminishes as μ/R_a^2. The valve should therefore have a high mutual conductance and a high internal resistance. This appears to be preferable to the use of an external resistance as a means of increasing R_a.

(iii) *High ratio of C to L.*—The advantage of making C as large as possible is also apparent from equation (23). (A large value of C is desirable on other grounds, referred to below.) As in (ii) there is a limit to the increase, set by the consequent decrease in the negative resistance term.

(iv) *Oscillatory circuit of low resistance.*—LLEWELLYN has pointed out[1] that in most circuits the resistances introduced by connection to the valve (*i.e.*, the resistive terms of groups (*b*) and (*c*) in equation (20) are so much greater than the actual resistance of the circuit, that the latter can be neglected in

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comparison. It should not be overlooked, however, that if means are adopted for minimising grid-current and the valve has a high value of R_a or an added external resistance, the coil resistance may become a limiting factor in the L/C ratio or in the added external resistance. On these grounds it is desirable that R shall be as small as possible and that the remainder of the circuit shall be designed in such a way that the added resistances due to the valve shall be kept as small as possible. It is seen from equation (23) that the conditions for this are the same as the conditions for minimising the reactance effects of the valve.

(v) *Use of stabilising reactance.*—The paper by Llewellyn is chiefly concerned with the elimination of the reactance effects of the valve by balancing the system by means of added reactances in the electrode connections. As applied to the case under consideration, this means finding some value for X_a such that the total of the reactance components of groups (*b*), (*c*) and (*d*) is zero. Under these conditions, the frequency of oscillation is the ideal value ω_0, and the relation

$$\omega_0^2 LC = 1$$

can be used in simplifying the analysis. It can be shown that the value

$$X_a = -\frac{1}{\omega_0 C} \cdot \frac{1 - k^2}{k^2};$$

where

$$k^2 = M^2/L\,L_g \quad \ldots\ldots\ldots\ldots (24)$$

will satisfy this condition. This means that X_a is the reactance of a condenser of magnitude

$$C_a = C \cdot \frac{k^2}{1 - k^2} \quad \ldots\ldots\ldots\ldots (25)$$

Thus C_a is proportional to C and their controls may be ganged if desired. Llewellyn's paper gives similar calculations for a great variety of circuits, and experimental results confirming that a very high degree of stability with respect to supply-voltages is realisable in this way.

In practice it would probably be desirable that by the means indicated in paragraphs (i)–(iv) above, the individual reactance effects should be minimised and then, since they cannot be eliminated in this way altogether, balanced by the appropriate external reactance.

(vi) *Stabilisation by close coupling.*—It will be noted that if $k = 1$ (*i.e.*, with unity coupling between the grid and anode windings), C_a becomes infinite. It can otherwise be shown that if $k = 1$. the reactance components of groups (*b*), (*c*) and (*d*) of equation (20) are zero for all values of R_a and R_g. LLEWELLYN gives experimental confirmation of this in circuits in which an effective coupling-coefficient of unity is obtained by special

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means, and shows also that the closer the coupling the less important is the correct adjustment of the balancing reactance. This result would appear to be at variance with the practically derived conclusions given in some of the abstracts—particularly 3 and 19, where a loose coupling is specifically recommended and apparently justified by measurement. The explanation probably lies in the fact that in the experimental work described in 3 and 19 it was not only k but M that was varied. The weight of the evidence is in favour of minimum M consistent with oscillation, obtained, not by loose coupling, but by tight coupling with small grid-inductance.

In this connection, it is suggested that toroidal coils wound on the new dust-core materials might prove to have advantages due to close coupling.

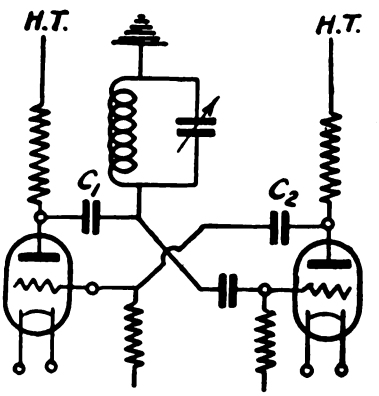

FIG. 3.

2.4. SPECIAL CIRCUIT-ARRANGEMENTS

The principles enunciated in the preceding sections are quite general in so far as they prescribe conditions under which the oscillatory circuit becomes the dominating factor in determining frequency, and where high degrees of stability are attained by special circuit arrangements, it is probable that, however complex the circuit operation may appear, it is, in fact, a means of satisfying the above fundamental requirements. Thus the arrangements described in abstracts 2.1, 2.2 and 2.3 are clearly of this kind.

The circuit used in the Franklin constant-frequency master-oscillator is illustrated in Fig. 3. (See abstracts 10.1 and 10.3.) This circuit makes clear the fundamental mechanism of self-maintenance of oscillation as described, for example, by HORTON (abstract 5.1). An e.m.f., assumed to be produced in the oscillatory circuit, is amplified by the second valve and "fed back" to the grid of the first valve with appropriate phase-reversal through the small coupling capacitance C_2. It is possible that a

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phase-controlling mechanism introduced in the back-coupling would lead to enhanced frequency-stability.

The oscillatory circuit can be regarded as " shunt connected " to the first valve, through the small coupling capacitance C_1. It could be inserted direct in the anode circuit, as in Fig. 4. If now the resistance load in the anode circuit of the second valve be replaced by a tuned circuit, we arrive at the arrangement

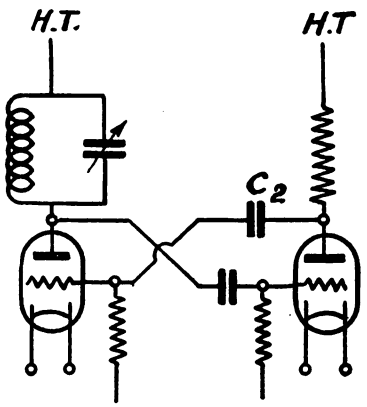

FIG. 4.

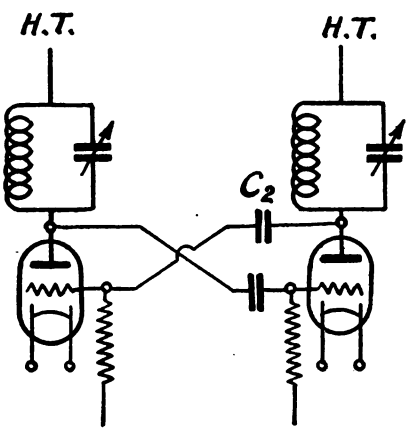

FIG. 5.

shown in Fig. 5, which is actually the Ross Gunn circuit described in abstract 7 (except that screen-grid valves are used in the latter). It is certainly an arrangement which may be expected to give the correct phase relationships for the maintaining voltages, and this is probably the reason for the high stability obtained. A rather special and unusual mode of interpretation is adopted in 7, but it appears from the above derivation

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that no new principle is, in fact, involved. The arrangement appears to be a very satisfactory one in practice, but it is not certain that equally good performance could not be obtained by simpler means.

The resistive stabilising arrangements described in abstract 17 seem to rest mainly on an empirical rather than a theoretical basis, but are none the less valuable, particularly as they would seem to be applicable to cases where substantial output-power is required.

With regard to the " long-line " circuits described in abstract 16, circuits which may prove to be of great value as far as short-wave operation is concerned, the difference between such circuits and those considered in the above general discussion lies in the use of distributed inductance and capacitance rather than concentrated inductance and capacitance. The full significance of this distinction from the practical point of view has not yet been explored analytically, but one important distinction, which will be referred to later in the section dealing with oscillations of complex wave-form, is that such a system is tuned not only to the fundamental, but to all the harmonics. This, as will be shown later, may be expected to minimise, if not to eliminate, a frequency-aberration due to the inevitable presence of harmonics in a self-maintained system. A further useful characteristic is that the effect on the natural sustained frequency of interelectrode capacitance and any variations in this, is likely to be very much less than in the normal capacitance-tuned circuit.

The chief advantages of the long-line schemes described would seem to be their direct applicability to transmitters or to high-powered master-oscillators, with a resulting stability sufficient for practical purposes.

The Dow circuit (abstract 8) is a case in which high stability is attained by methods not obviously related to the general principles enunciated above. Here, however, the inherent stability with respect to operating conditions is not notably greater than that obtainable by the general methods indicated, and appears to be due in the main to a mutual compensation of contrary variations arising from supply-voltage changes. The outstanding feature of the Dow circuit would seem to be rather the special method of transfer of potential to the load-circuit by way of the additional valve-electrode. This is considered more fully in section 6.

The arrangement described in abstract 23 is of special interest, and may be expected to lead to further useful developments. In the form described it is similar in principle to the automatic grid-bias controlled mechanism of the usual grid-leak and condenser circuit, but the grid-bias control is obtained from a separate rectifier-valve inductively coupled to the oscillatory circuit.

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This allocation of two distinct functions (oscillation and grid-bias control) to two separate valves will give a new element of elasticity to the design of the oscillator, and permits of the choice of a rectifying mechanism specifically adapted to control of the amplitude of the generated oscillation. It is claimed that a greatly improved linearity of operation (*i.e.*, an improved linearity of the dependence of amplitude on an applied grid-bias or modulating voltage in the grid of the first valve) is obtainable in this way, and the claim is substantiated by oscillograph records. It is also claimed that improved frequency-stability is obtained, but no data are given on this point, though the arrangement seems likely to lead to a reduction of " scintillation " in modulation as a consequence of the improved linearity of control.

2.5. Conclusions relating to Methods of Stabilisation Derived from the Simplified Analysis

The simple and admittedly inadequate analysis of the valve-maintained oscillator given in this section indicates that frequency variation due to incidental changes in the valve-conditions can be minimised by eliminating grid-current as far as possible, by using valves of high mutual conductance and high internal resistance or by adding a high resistance in the connection to the valve, by using circuits of low resistance with a high ratio of inductance to capacitance, by close coupling between grid and anode windings with low grid circuit inductance, and by connecting suitable balancing impedances in the electrode connections, or by using special circuit arrangements such as those described in abstracts 7, 16 and 17. It appears that by such means master-oscillator stabilities can be obtained which are variously estimated at between one and a hundred parts in a million, as far as changes due to valve-conditions alone are concerned. There are, however, other factors involved which are of greater effect and which will be referred to later.

Not all of the means indicated above are applicable to transmitters or to cases where appreciable power has to be supplied by the oscillators, but even in such cases they are advantageous in so far as they can, in fact, be applied, and abstract 19 gives information of a case in which an improvement up to a stability of a few parts in 10,000 was obtained in a transmitter, simply by attention to these points of detail in design.

3. MORE EXACT THEORY OF VALVE-MAINTAINED OSCILLATION

3.1. The Nature of the Problem

For an exact theory of valve-maintained oscillation, it is necessary to take into account the non-rectilinearity of the characteristic of

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the valve concerned. This introduces formidable analytical difficulties, particularly as the valve-characteristics do not lend themselves to representation in simple functional form. This part of the theory is, in consequence, comparatively unexplored, though a number of valuable contributions are listed in the abstracts (Nos. 11.1 to 11.3, 12.1 to 12.5).

It appears that the problem is ultimately resolvable into the solution of the differential equation

$$\ddot{x} + f(x)\dot{x} + \omega_0^2 = 0 \quad \ldots\ldots\ldots\ldots (26)$$

where $f(x)$ is a polynomial in x. Unfortunately, no explicit and general solution can be found, and the papers listed above are concerned with different types of approximate solution. APPLETON, VAN DER POL, ROCARD and GIOVANNI have given solutions in terms of successive approximation, with a restricted power range for $f(x)$. VAN DER POL, LIENARD, HORTON and USUI have described what are essentially graphical methods of solution. It seems desirable that both these methods should be further explored, with emphasis on physical interpretation and practical application.

It is further necessary to point out as a factor which somewhat limits the practical value of the work already carried out on these lines, that the analysis has not so far been applied to the oscillating system as a complete network including the interelectrode capacitances, which are likely to play an important rôle at high frequencies.

When all the factors concerned are taken into account, the complexity of the problem from an analytical point of view makes it unlikely that an accurate quantitative solution of general application will ever be available in a manageable and practicable form. There is, however, the possibility of obtaining information of a qualitative kind which would prove useful in design and a continued exploration on these lines is therefore advocated.

A great deal of practical importance attaches to the question of harmonics and their effect on frequency. It is immediately obvious that, apart from the possibility suggested in the last paragraph of section 2.4, the generation of pure sine-wave oscillation by means of simple valve-circuits is inherently impossible. This is due to the fact that the indefinite amplitude deduced from the simple linear theory is inconsistent with the physics of the case, and the only factor which determines and limits amplitude is a departure from rectilinearity of some element of the system. This departure is inseparable from the production of harmonics, so that finite and limited amplitude is itself inseparable from the production of harmonics. The significance of this from the point of view of frequency-stability is that the production of harmonics, to an extent depending on amplitude, is associated, necessarily it seems, with an aberration of the fundamental frequency.

The first of these points is very clearly brought out by ROCARD (abstract 12.4). He gives, for example, as the amplitude of the

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third harmonic (A_3) in terms of that of the fundamental (A_1) for the approximate solution for a " back-coupled " oscillation of restricted amplitude

$$A_3 = -\frac{RC\omega R_a + \omega L + \mu M\omega}{8R_a} A_1 \quad \ldots\ldots (27)$$

He points out that

$$RC\omega R_a + \omega L + \mu M\omega < O$$

(M presumably to be taken as negative) is the limiting condition of maintenance, and the third-harmonic content is a measure of the departure from this limiting condition.

The second point is also brought out by Rocard, who gives as the frequency of maintenance under these conditions

$$\omega = \frac{1}{\sqrt{LC}} \sqrt{1 + \frac{R}{R_a} - \left(\frac{4A_3}{A_1}\right)^2} \quad \ldots\ldots (28)$$

which may be compared with the limiting value for zero amplitude.

$$\omega = \frac{1}{\sqrt{LC}} \sqrt{1 + \frac{R}{R_a}} \quad \ldots\ldots (29)$$

3.2. Analysis Based on Energy-Changes

The analysis contemplated in 3.1 is in terms of amplitude, and is the only method of getting complete information with regard to amplitudes and phases of individual components of the total oscillation. There is another possibility, however, which may prove of great value from a practical point of view, even though it cannot, theoretically, give as completely detailed a solution as the former method. This alternative is an analysis of the energy-changes within the system. The idea is very briefly outlined by Groszkowski (abstract 13), who writes as follows :—

> " Equations based only on the fundamental frequency do not represent sufficiently the energy-exchanges of the system ; but only these changes can be taken as a basis in considering a system whose frequency is determined by conditions in which free oscillations of electric charge are occurring. . . . It can be proved that the frequency-variation is strictly related to the content of harmonics. . . . The imaginary energy which corresponds to the oscillation of the electric charge between the capacity C and the inductance L . . . must be balanced within the circuit itself. . . . On this assumption one can obtain the relation

$$-\frac{\delta f}{f} \propto D^2 \quad \ldots\ldots (30)$$

where $D = \sqrt{\dfrac{\text{sum of squares of harmonic amplitude}}{\text{square of fundamental amplitude}}}$ . . (31)

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and f = critical or zero-amplitude frequency.

. . . Physically, the influence of the content of harmonics on the frequency-variation can be explained as follows: If the oscillations are purely sinusoidal the fundamental frequency . . . is determined by*

$$f_0 = \frac{1}{2\pi\sqrt{LC}}$$

and therefore the energy-distribution in both branches is equal. When the harmonics appear, the currents corresponding to them are flowing chiefly through the condenser branch, and therefore they increase the energy in this branch in comparison with the branch LR.

In order to keep the energy-equality in both branches, the fundamental frequency must slightly diminish itself in respect to f_0. The current in the branch LR increases then slightly in order to allow its electromagnetic energy to increase suitably. It seems that this principle is quite general. It applies to all oscillating systems in which the frequency is determined by the constants of the system."

GROSZKOWSKI has not yet published the full development of this theme.† The above is taken from a preliminary note in the form of a letter published in the *Wireless Engineer*. It is clear, however, that the idea should prove of great value in the analysis of complex oscillation. Actually, the form in which the principle is enunciated above does not seem sufficiently generalised. GROSZKOWSKI has assumed that the circuit itself will show a reactive power balance, but the valve-capacitances will also be the seat of reactive power changes, and a more general condition is balance of reactive power and of real power in the complete system. Physically, this states that energy consumed in one part must be supplied in some other part, and that the mean value of the kinetic (magnetic) energy is equal to the mean value of the potential (electric) energy. The first gives a maintenance-condition and the second a frequency- and wave-form condition.

It may be noted that Groszkowski's general result (equations (30) and (31)) given above is consistent with the particular result given by Rocard, equations (28) and (29).

* This is assuming equal resistances in the inductive and capacitive branches of a dynatron circuit. (See 2.3.1.)

† Since the above was written, an English version of Groszkowski's paper has appeared in the P.I.R.E., vol. 21 (July 1933), pp. 958–981, under the title "The Interdependence of Frequency-Variation and Harmonic Content and the Problem of Constant-Frequency Oscillation."

A paper by E. B. Moullin on the same subject has also been published recently in the J.I.E.E. (vol. 73, pp. 186–195) under the title "The Effect of the Curvature of the characteristic on the Frequency of the Dynatron Generator."

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3.3. Application of Energy-Analysis to Dynatron Circuit

It will be of interest to show the application of the analysis based on energy exchanges to the simple dynatron circuit considered in section 2.3.1, and illustrated in Fig. 6. For simplicity of illustration, the driving impedance will be considered as a pure negative resistance.

Considering first the limiting case of pure sine-wave oscillation, *i.e.*,

$$v = V \sin \omega t \quad \ldots\ldots\ldots\ldots (32)$$

the reactive-power balance is given by

$$\frac{1}{2} LI^2 = \frac{1}{2} CV^2 \quad \ldots\ldots\ldots\ldots (33)$$

Also

$$I^2 = \frac{V^2}{R^2 + \omega^2 L^2} \quad \ldots\ldots\ldots\ldots (34)$$

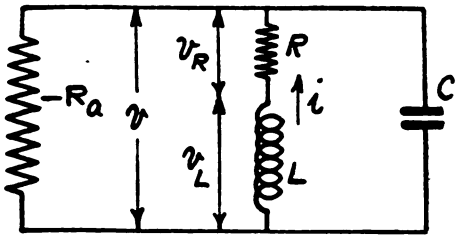

Fig. 6.

Therefore

$$\frac{LV^2}{R^2 + \omega^2 L^2} = CV^2 \quad \ldots\ldots\ldots\ldots (35)$$

Whence

$$\omega^2 = \frac{1}{LC}\left(1 - \frac{CR^2}{L}\right) \quad \ldots\ldots\ldots\ldots (36)$$

as in equation (12) of section 2.3.1.

For the real-power balance

$$\frac{V_R^2}{V^2} = \frac{R^2}{R^2 + \omega^2 L^2} \quad \ldots\ldots\ldots\ldots (37)$$

and the mean power consumed in R is given by

$$\frac{V_R^2}{R} = \frac{R}{R^2 + \omega^2 L^2} V^2 = \frac{CR}{L} V^2 \quad \ldots\ldots\ldots\ldots (38)$$

in virtue of equation (35). Equating this to the power supplied by the negative resistance, we have

$$\frac{V^2}{R_a} = \frac{CR}{L} V^2 \quad \ldots\ldots\ldots\ldots (39)$$

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or

$$R_a = \frac{L}{CR} \quad \ldots\ldots\ldots\ldots (40)$$

which is the maintenance-condition given in equation (11) of section 2.3.1.

Coming now to the actual case, where

$$v = \Sigma\, V_n \sin n\omega' t \qquad n = 1, 2, 3, \text{etc.} \quad \ldots\ldots (41)$$

ω' being the new fundamental frequency, the reactive-power balance gives

$$L\Sigma \frac{V_n^2}{R^2 + n^2\omega'^2 L^2} = C\,\Sigma_1\, V_n^2 \quad \ldots\ldots\ldots (42)$$

The harmonic terms on the left-hand side decrease very rapidly owing to the presence of n^2 in the denominator, and because V_n^2 is inversely as n^2 on account of the parallel capacitive branch. The sum therefore will differ very little from the value of fundamental term only. Thus approximately

$$L \frac{V_1^2}{R^2 + \omega'^2 L^2} \simeq C\Sigma V_n^2 \quad \ldots\ldots\ldots (43)$$

Putting

$$\omega' = \omega + \delta\omega \quad \ldots\ldots\ldots\ldots (44)$$

$$\frac{LV_1^2}{R^2 + \omega^2 L^2 + 2\omega\delta\omega L^2} = CV_1^2 + C\,\Sigma V_n^2 \qquad n = 2, 3, 4, \text{etc.} \quad (45)$$

Since

$$R^2 + \omega^2 L^2 = L/C \text{ from equation (35)},$$

$$\frac{LV_1^2}{\frac{L}{C} + 2\omega\delta\omega L^2} \simeq CV_1^2 + C\,\Sigma V_n^2, \qquad n = 2, 3, 4, \text{etc.} \quad (46)$$

i.e.,

$$V_1^2 (1 - 2\omega\delta\omega LC) \simeq V_1^2 + \Sigma V_n^2 \quad \ldots\ldots (47)$$

and since

$$\omega^2 = \frac{1}{LC}\left(\frac{1 - CR^2}{R}\right) \text{ from equation (36)}$$

we have finally

$$-\frac{\delta\omega}{\omega} = \frac{1}{2} \frac{1}{1 - \frac{CR^2}{L}} \cdot \frac{\Sigma V_n^2}{V_1^2}; \qquad n = 2, 3, 4, \text{etc.} \quad (48)$$

which is the particular form of Groszkowski's general result (30 & 31) for this case.

The real power balance leads to

$$\frac{CR}{L}\Sigma V_n^2 = \frac{1}{T}\int_0^T \frac{V^2}{R_a}\, dt \quad \ldots\ldots\ldots (49)$$

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which cannot be developed further without some assumption as to the nature of the variation of R_a with V.

The reactive-power balance equation (48) already gives useful design-information. It shows, for example, the value of making C large, not on account of the term CR^2/L, which is very small in any case, but on account of the consequent reduction of ΣV_n^2. It further suggests that the limiting curvature should be such as to produce high order harmonics in preference to low, since V_n^2 is inversely as n^2. Apparently, therefore, a sensibly rectilinear characteristic with considerable and abrupt curvature at its limits should show less frequency-aberration than a uniformly curved characteristic. This is a matter for investigation.*

3.4. "Long-line" Controlled Systems and Frequency Aberration due to Harmonics

The above discussion of the effect of harmonics on the fundamental frequency may throw some light on the operation of the "long-line" circuits described in abstract 16. From the point of view of harmonics, the transmission-line differs from the normal tuned circuit in that its impedance is non-reactive not only to the fundamental but also to all the harmonics (assuming a reasonably low attenuation factor). It may therefore be expected that such a circuit, maintained in oscillation by a valve, should be practically free from the frequency-aberration due to harmonics.

The subject requires a detailed analysis before the full practical significance of this distinction can be assessed, but the above is suggested as a possible reason why at least one factor in frequency-instability may be expected to be eliminated in such systems.

3.5. Oscillation Maintained by Impulsive Energy-Supply

Edgeworth has pointed out (abstract 2.3) that if a pendulum be maintained in oscillation by impulses of infinitesimal duration delivered at the moment of maximum velocity (zero displacement) the frequency should not be affected by the maintenance. The electrical analogy is an impulsive supply of energy to the oscillatory circuit, of very short duration, phased to correspond to the current-maximum in the circuit.

The possibility has not yet been fully explored. It is unlikely that it would be found applicable to self-maintained transmitting systems, owing to the practical difficulty of supplying the energy in a pulse of sufficiently short duration. It might, however, prove practicable in the case of small-power master-oscillators. In

* See also Groszkowski's paper in P.I.R.E., July 1933, referred to in footnote on p. 15, for further practical applications and experimental verification of his analysis of the effect of harmonics.

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particular, it might be found suitable for the maintenance of tuning-forks or similar mechanical frequency-standards. The Thyratron or similar gas-filled relay-valve may perhaps find application in this way.

The multivibrator type of circuit can be made to give an approximately rectangular wave-form. The differential of this would give the impulsive type of excitation required.

Transmitters have been designed in such a way that anode current only flows for a part of the radio-frequency period, but the objective in this case has been increased d.c.-a.c. power-efficiency. Abstract 21 describes what appears to be the only application of this principle to a small-power oscillator, specifically with a view to frequency-stability. It is claimed to give stability to 5 parts in a million for 15 per cent. changes in supply-voltages, but very little detail is given.

3.6. Conclusions relating to the More Exact Theory of Valve-Maintained Oscillation

It is apparent that, apart from the possibility referred to in section 2.4, valve-maintenance of oscillation is inseparable from the production of harmonics with a consequent effect on fundamental frequency. Experimental data are lacking as to the quantitative significance of this effect, but Groskowski's formula suggests that a 1 per cent. R.M.S. harmonic content (in the voltage across the tuning capacity) would give a frequency-change of the order of 50 parts in a million, from that corresponding to zero amplitude. This instability of frequency would arise as a consequence of variation in amplitude and could thus be minimised by maintaining constancy of amplitude as far as possible. (See last paragraph of section 2.4.)

The effect can also be minimised by means of circuits designed to reduce as far as possible the potential differences due to harmonics, though it is unlikely that the use of anything more complicated than the maximum permissible tuning capacity, or perhaps a small resistance in the capacitance branch, would be justified by the consequent improvement in frequency stability.

There is, moreover, another effect likely to arise from variation in amplitude which will, in general, outweigh the harmonic effect in significance. This is considered in the next section.

4. EFFECT OF VALVE INTERELECTRODE-CAPACITANCE ON FREQUENCY

In so far as the valve interelectrode-capacitances can be considered as constants, their effect on frequency is immediately derivable from the analyses already given, and they would not,

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moreover, figure in themselves as significant factors in frequency-instability. From the nature of these capacitances, however, it is improbable that they can be regarded as constants in the above sense, for their effective magnitude may be expected to vary with the space-charge conditions of the valve, which, in turn, will vary with supply-voltage and oscillating conditions.

A brief investigation of the frequency-stability of the dynatron circuit carried out by the present writer in 1930 showed frequency-variations with supply-voltage which were in some cases more than ten times as large as the simple theory of the circuit would have led one to expect. It was suggested at that time in a report to the Radio Research Board that a variation of interelectrode (anode-screen grid) capacitance with oscillating conditions was a possible cause, and some experimental evidence consistent with this view was given. The experiments were not quite satisfying, however, as no separation of the harmonic effect on frequency was attempted. To account for the observed frequency-aberrations in terms of the harmonic effect alone would, however, require something like a 6 per cent. *change* in harmonic content with a 20 per cent. change in anode-voltage. This is not impossible, but is very improbable.

HARTSHORN (abstract 22) has made a brief investigation of the variation of interelectrode-capacitance with space-charge, but there is at present an insufficiency of experimental information on the point.

An experimental investigation of the matter is now being carried out at the National Physical Laboratory. The preliminary results so far obtained, and also some similar results found by MOULLIN and now in course of publication*, show that variations in inter-electrode-capacitance with conditions of operation may be of the order of several hundredths of a micro-micro-farad, and may therefore be a very significant factor in frequency-variation—possibly even the most important factor in some cases, where the inter-electrode-capacitance is a direct addition to the tuning capacitance. The existence of this effect does not necessarily invalidate the various circuit-arrangements hitherto considered, in which it was not explicitly taken into account, but it implies that where frequency-stability is obtained by the empirical adjustment of some circuit-element (*cf.* abstracts 1 and 2.4), the effect of the latter is to compensate for the resultant of all the factors causing frequency-variation, the interelectrode-capacitance effects being included. This point is referred to explicitly in section 2.3.1.

A large tuning capacitance is clearly desirable as a means of minimising the effect of any change in interelectrode-capacitance. A large resistance in series with the connections to the electrodes is

* See "The Interelectrode Capacitance of the Dynatron, with special reference to the Frequency-Stability of the Dynatron Generator." G. B. Baker. See also the discussion on this paper. (J.I.E.E., vol. 73, 1933, pp. 196–203.)

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a further expedient, but one which is limited by oscillation-maintenance requirements. A further possibility is to ensure constancy of amplitude as far as this may be practicable, and to arrange the supply-and valve-conditions in such a way that the space-charge distribution (as judged, for example, by the mean anode-current) is not materially dependent on the oscillation-amplitude. Thus, for example, a battery grid-bias might prove superior to automatic or grid-leak controlled bias on this account.

On the practical aspect of the matter, however, very little can yet be said owing to the lack of data. It requires specific investigation.

5. THE DEPENDENCE OF OSCILLATION-FREQUENCY ON CHANGES IN THE PHYSICAL CONFIGURATION OF THE SYSTEM

5.1. Changes in the Oscillatory Circuit

5.1.1. *Secular changes*

In so far as, by careful design, the self-maintained frequency of oscillation is made to depend chiefly on the LC product for the oscillatory circuit, the constancy of this product or of its individual terms is clearly the limiting factor in stability.

Changes in the LC product may be classified into two main groups, which may be described as temporary and secular respectively.

The second includes all gradual and, in general, non-reversible changes in physical configuration due to corresponding slow changes in the materials involved. Their minimisation, in so far as this is practicable or desirable in any given case, is part of the general problem of robust and stable construction with materials of known stability. There is at present very little information about the secular stability of the synthetic insulating materials commonly employed in the construction of coils and condensers, and, in the absence of such information, all that can be said is that any construction employing such materials, *e.g.*, ebonite, keramot, bakelite, mycalex, etc., cannot be expected to show a high degree of secular stability and may be very defective in this respect, considered in relation to the very severe requirements of frequency-stability. Pure metals, on the other hand, or well-annealed alloys of certain kinds, such as brass and, as insulation, quartz, fused quartz, or marble, may be expected to be stable (though recently changes of appreciable amount, *i.e.*, 4 to 10 parts in 100,000 in 6 years, have been recorded in marble cylinder coil supports).

Such secular changes are, however, less important from a practical point of view (except in relation to standard apparatus)

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than the temporary changes due to environment and conditions of operation.

The most important factor is temperature, though atmospheric pressure is also significant.

5.1.2. *Variations due to temperature and their minimisation*

The actual degree of the temperature-effect will clearly depend on a great variety of constructional details, and no representative temperature-coefficient can be given. It appears, however, that in most cases of master-oscillator control, temperature is likely to be the most important single factor. For example, in experiments carried out at the National Physical Laboratory, H. A. THOMAS has found that all effects other than temperature can be brought within the order of parts in 100,000, whereas temperature effects may amount to as much as 1 in 10,000 per degree Centigrade. Similar effects may be expected in self-maintained transmitters, but in these cases there may be other factors (*e.g.*, effect of aerial variations) of comparable significance.

In general, the inductance of solenoidal coils of ordinary construction increases with temperature, for though there will be an increase both in length (*i.e.*, spacing) and diameter, the latter effect, with a consequent increase in inductance, predominates. In fact, for a self-supported coil, the theoretical temperature-coefficient of inductance, neglecting possible effects of temperature on the elastic properties of the strained material, will be the linear expansion-coefficient of the material of the coil. The extent of the variation may be expected to amount to parts in 100,000/°C.

W. H. GRIFFITHS, utilising the expansions of the materials of the coil supports, has designed a construction for which self-compensation is claimed in coils of efficient shape. This is described in abstract 20.

As between master-oscillators and self-excited power-systems, there is obviously an important difference in thermal conditions. The small power-dissipation in the first case makes environmental temperature the chief factor, and permits of thermostatic temperature-control as a means of stabilising. Such a system is described in abstract 9. Alternatively, a comparatively simple self-compensated construction, such as that described by Griffiths (abstract 20) would seem to be applicable.

In self-excited transmitters, on the other hand, the coil temperature will be partly, and in some cases mainly, dependent on the energy-dissipation in the coil (*i.e.*, of course, the I^2R loss in the coil itself). The change in inductance due to this cause can be minimised by keeping the effective conductor cross-section large and of large surface to give effective air-cooling. In the case described in abstract 19, an initial frequency-drift of 7 parts in 10,000 in 2 hours was reduced to one of 5 parts in 100,000 simply

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by substituting for the original coil one constructed of copper tube about 2 cm. in diameter.

A further alternative is to employ a coil of special construction in which changes due to increase of temperature are so constrained as to cause a minimum of change in inductance. A coil of this kind is described by URE, GRAINGER, and CANTELO (H.M. Signal School, Portsmouth) in their paper "The balancing and stabilizing of high-frequency Amplifiers." (J.I.E.E., vol. 72, pp. 528–543.)

For master-oscillator systems without temperature-control, there are two main alternatives—stabilisation of LC or stabilisation of L and C separately. The Griffiths construction already described is an example of stabilisation of L. Abstract No. 10 describes the most recent Marconi master-oscillator system, where the residual temperature-coefficient of a plain solenoidal coil wound on a loaded ebonite former (dimensioned so as to give partial thermal compensation), and the temperature-coefficient of the associated small variable condenser, are balanced by a small change in an auxiliary capacitance which is determined by the differential expansion of a brass rod and the coil former.

Very little has so far been done in the design of temperature-stabilised variable condensers. This subject is being explored by H. A. Thomas at the National Physical Laboratory. In the absence of special design for thermal stability the variation with temperature may be expected to be at least between 1 and 10 parts in 100,000/°C.

ROSS GUNN (abstract 7) has made use of small auxiliary condensers having one plate bimetallic, as a means of obtaining an overall compensation for ambient-temperature changes. This is one of many possible applications of bimetallic thermal control to overall compensation.

5.1.3. *Variations due to atmospheric pressure and humidity*

Where air-dielectric condensers are used, it appears that variations of capacitance due to variation of the dielectric constant of air with pressure may be significant if a very high degree of frequency-stability is required. H. A. THOMAS finds that a variation amounting to about 20 parts in 10^6 (in capacitance) may be expected for a change in pressure equal to one inch of mercury.

There are no direct data as to the effect of humidity. The dielectric constant of steam at atmospheric pressure over the range 140°C. to 150°C. is given in the International Critical Tables as $1{\cdot}00785 - 0{\cdot}00016\ (t - 140°)$, and that of air at 19° as 1·000576. From these data, and the known values of the vapour pressure of water, it appears that at ordinary temperatures the capacitance of an air-dielectric condenser might depend on the humidity to the extent of a few parts in 100,000 at least.

The stable condenser at present being designed and constructed at the National Physical Laboratory is hermetically sealed with a view to eliminating the effects of changes of pressure and humidity.

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5.2. *Changes in the Valve*

Apart from the above physical changes in the oscillatory circuit, physical changes in the maintaining valve may also contribute to frequency-variation, these changes being due either to the direct effect of ambient temperature or to the establishment of thermal equilibrium in the valve itself.

Assuming a system designed for minimisation of the normal effects of variation of maintenance-conditions on the frequency, the valve interelectrode-capacitance will still, in general, remain a significant factor, and any variation of this (and also to a less extent, any variation of the valve-characteristics as a whole) with ambient temperature will cause frequency-variation. There is at present no information as to the magnitude of such effects in transmitting valves, and very little relating to small valves such as might be used in master-oscillators. A preliminary investigation by Thomas, at the National Physical Laboratory, suggests that in the frequency-variations with temperature of the circuit employed in his measurements, the effect of ambient temperature on the valve was negligible. This one might anticipate as a consequence of the steep temperature-gradient in the valve itself due to the heated filament. On the other hand, it would be expected that there would in any case be an initial period of drift while thermal equilibrium in the valve was being established, though the results obtained by Aguieff and Kopytin (abstract 19) seem to imply that this effect is very small.

On these points there is need for further specific investigation.

6. FREQUENCY-VARIATIONS DUE TO UTILISATION OF OSCILLATORS

In addition to variations due to causes of a purely electrical character associated with the maintenance of oscillation, and variations due to changes of a physical nature, mainly due to environmental changes, the utilisation of an oscillator for the excitation of a transmitting aerial or for the excitation of an amplifier will introduce further possibilities of frequency-instability.

If the coupling of an oscillator to the utilisation-circuits had no effect other than increase of effective resistance in the oscillatory circuit, the means of frequency-stabilisation discussed in the preceding sections should minimise frequency-variations due to the utilisation of the oscillator, for they aim at the realisation of conditions in which the oscillation-frequency is independent of the effective resistance of the oscillatory circuit. Again, even when the coupling to the utilisation circuits introduces reactance into the oscillatory circuit, the conditions for stability are not violated as long as the reactance is constant in value. In practice, however, whether the utilisation be the supply of voltage to an amplifier or

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of power to an aerial, it is very difficult to satisfy the above conditions, particularly when a variation of load or coupling has to be provided for. Dow (abstract 8) gives a statement on this point which may be considered as generally applicable to the utilisation of master-oscillators. " Whenever capacitative, inductive, or direct coupling was employed between known forms of oscillators and any terminal apparatus, changes in the latter were found to influence the frequency of oscillation so long as this coupling existed to such an extent as permitted of appreciable amounts of energy to be derived from the oscillator. This influence of terminal apparatus upon frequency could, of course, be reduced appreciably by very loose coupling or by coupling to derive energy from the oscillator at some harmonic of the fundamental frequency " (*cf.* abstract 9),

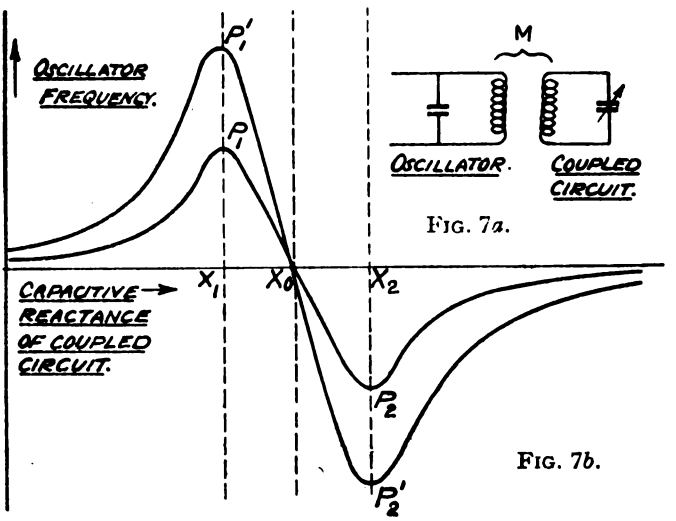

FIG. 7*a*.

FIG. 7*b*.

FIGS. 7A AND 7B.

" but even under these conditions the effect of terminal apparatus was found to be quite noticeable, especially since it was desired to keep all possible accumulative sources of error to within 0·015 per cent."

The above refers to master-oscillator systems, but is obviously applicable, with even greater force, to cases where transfer of appreciable power to the utilisation-circuit (*e.g.*, an aerial) is a condition of operation.

In view of the wide variety of practical conditions, a generalised treatment of the subject is of limited utility, but the following is a point of some practical interest.

It is assumed that a tuned circuit is coupled to an oscillator, as shown in Fig. 7 (*a*). It is well known that as the secondary-circuit

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tuning capacitance is varied through resonance the oscillator-frequency is varied as shown by the curves $P_1 P_2$ and $P'_1 P'_2$ in Fig. 7 (*b*) (the latter corresponding to a closer coupling than the former).

The point X_0 corresponds to resonance of the secondary circuit with the original oscillator-frequency. It is a condition corresponding to maximum transfer of energy from the primary to the secondary circuit, and, moreover, it is a condition in which the effective reactance, and therefore the oscillation-frequency, of the primary circuit is not affected by the coupling to the secondary circuit. This would, therefore, seem to be a suitable condition of coupling, and is, in fact, a condition which is often used in practice. It should be noted, however, that it is the worst possible condition from the point of view of frequency-stability with respect to small variations of the reactance of the secondary circuit.

Over the whole range of possible values of secondary-circuit reactance there are only two regions where the oscillator-frequency has a stationary value with respect to secondary-circuit reactance, namely, those corresponding to the points X_1 and X_2. These will, of course, correspond to a diminished transfer of energy as compared with the resonant condition X_0, but in practice it will probably be possible to compensate for this by a closer coupling. Further, if it is desired to operate on the original undisturbed oscillator-frequency, the tuning condenser of the oscillator can be readjusted so as to produce this condition.

The general point is that the coupling condition which corresponds to an undisturbed oscillator-frequency and the maximum transfer of energy for a given degree of coupling is likely to be, in practice, a particularly unfavourable condition from the point of view of frequency-stability and, where practicable, a stationary-frequency condition rather than an undisturbed-frequency condition should be preferable in this respect.

Such considerations may be of great practical significance in cases where an oscillator is inductively coupled to an aerial, the effective characteristics of which are liable to vary considerably with environmental conditions and may be subject, for example, to fluctuations due to movement by wind.

No reference has been found in the literature to this aspect of coupling in relation to frequency-stability, and it is not known whether the stationary condition is actually used in practice. It would seem desirable that its possibilities should be examined in relation to practical cases where the load-circuit reactance is subject to variation.

In present practice the effect of load on the frequency of master-oscillator systems is minimised by using a coupling as loose as practical requirements will permit, and by careful " neutralising " of amplifying circuits or the use of screen-grid valves in order to reduce input-impedance variations, and by the use of " buffer valve " circuits between the oscillator and amplifier.

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A new alternative is described by Dow (abstract 8) in which coupling is effected by variations of anode-current in a screen-grid valve of which the control and screen-grid only are the seat of self-maintained oscillating potentials. Given satisfactory elimination of capacitive coupling, either by "neutralising" or the use of an additional screening electrode (*i.e.*, a pentode valve) such a system should give an ideal uni-directional coupling, inasmuch as the oscillator-potentials are transferred to a load circuit but variations of load potential are not transferred back to the oscillator. The idea is comparatively new, but has already been found to give a very satisfactory elimination of frequency-variations due to the load.

7. AUTOMATIC FREQUENCY-MONITORING AS A MEANS OF SECURING FREQUENCY-STABILITY IN TRANSMITTERS

From the foregoing sections it is apparent that frequency-instability may arise from a great variety of causes, some of which cannot be eliminated in practical conditions of operation. This is particularly true of simple self-maintained transmitters, where the aerial characteristics have an appreciable effect on frequency.

Considerable practical interest therefore attaches to the possibility of the automatic monitoring of the frequency of a self-maintained transmitter by direct reference to some element of substantially constant frequency, for example, a small-power valve-oscillator designed for a high degree of frequency-stability under no-load conditions. In fact, this appears at present to be the only means available for securing the desired end in systems where elaboration of equipment is impracticable. It would have, moreover, the outstanding advantage of eliminating all frequency-variations, of whatever origin, within the limits of constancy of the auxiliary controlling element and subject to the restoring-force to frequency-change ratio of the controlling mechanism. It also embodies the possibility of controlling a number of transmissions on various frequencies by reference to a single source of stable frequency.

It may be well to emphasise at this point the distinction between master-oscillator and quartz-crystal or tuning-fork controlled systems, on the one hand, and the type of self-monitored system here contemplated on the other.

In the former, the finally emitted frequency is derived directly from the controlling and energising frequency, and is, in fact, equal to or some integral multiple of this frequency. The practical disadvantages of this method are the elaboration of equipment necessarily involved in the amplifying and frequency-multiplying

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stages required, and the difficulty of changing the frequency, either in steps, or uniformly in some permitted band.

In the contemplated monitored system, the emitted frequency is that of a self-maintained transmitter. This is linked up with the frequency of the controlling element in such a way that any deviation from the desired value automatically gives rise to a restoring force which tends to eliminate the deviation. The emitted frequency is not, therefore, in general, exactly equal to that of the controlling element, but will differ from it by an amount depending on the ratio of the restoring-force to frequency-displacement ratio of the mechanism, and to the time-lag of the controlling mechanism.

Considering the matter in more practical detail, it is clear that the requirement is some auxiliary controlling element giving an output-frequency controlling action which depends both in magnitude and sense on the frequency-change from some assigned mean value, and which, if possible, shall also be independent, in sense at least, of the amplitude of the oscillation of the transmitter.

The possibilities of this idea are still comparatively unexplored and its development will clearly depend on ingenuity and inspiration rather than on systematic research. Some existing possibilities may, however, be briefly reviewed.

Considering first the actual frequency-controlling mechanism in the transmitter itself, two alternatives are available, which may be classified as mechanical and thermionic respectively.

In the first, the emitted frequency is changed by a controlled change in the actual LC value of the oscillatory circuit. Abstract 17 describes a scheme in which the capacitance is varied by controlled changes in an auxiliary tuning-capacitance. Abstract 18.1 describes a system recently developed in France, in which L is varied by means of the frequency-change controlled flow of direct current through an auxiliary winding on an iron core which is also linked with the main inductance of the transmitter.

In the thermionic type of control a number of possibilities are at present available. One is to design the oscillator for frequency-instability with respect to supply-voltages and to apply the restoring force as a voltage to one of the electrodes, probably the grid. Another alternative is to associate with the oscillatory circuit an auxiliary valve so connected to the circuit that it acts as a pure reactance of amount depending on electrode-voltages, the latter being made dependent on the frequency-change. A system of this kind is described in abstract No. 17. A further possibility of similar character is described in abstract No. 18.2.

Comparing the thermionic and the mechanical mechanisms, the former would have the advantage in respect of inertia, but the latter is likely to permit of a much higher ratio of restoring-force to frequency-displacement.

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Considering now the conversion of the frequency-displacement into a restoring force, there are again two main alternatives at present available. The first is the use of a quiescent, as distinct from an oscillatory circuit, the steep slope of the resonance-curve being employed (*cf.* abstract 17, in which a quartz-crystal resonator is used), and the second is to refer the transmitter-frequency to that of a low-power master-oscillator (which may be quartz-controlled if desired) (*cf.* abstracts 17 and 18). In this case an audible heterodyne-frequency is produced and an audio-frequency discriminating element is also required, which may, for example, take the form of an audio-frequency resonant circuit operated off resonance. Any form of audio-frequency amplifier giving a response proportional to frequency will meet the requirement.

Comparing the above alternatives, the quiescent circuit has advantages in respect of simplicity, but the oscillating circuit and heterodyne method is likely to permit of a higher ratio of restoring-force to frequency-displacement.

As already pointed out, it is desirable that the sense of the restoring force shall depend only on frequency-displacement and not on oscillation-amplitude. This condition can be satisfied by making the restoring force depend on the difference between the responses of two auxiliary elements, one of which is proportional to amplitude only and the other to amplitude and frequency-displacement. These can be realised, for example, by an aperiodic or untuned coupling, and a sharply-tuned coupling, combined in opposition. Alternatively, the opposed rectified responses of two oscillatory circuits, tuned respectively a little above and a little below the desired frequency of operation, would give a nett response of which the sense was quite independent of amplitude and dependent only on the sense of the frequency-displacement (though its intensity would depend on amplitude). The same principle would be equally applicable to frequency-control by the heterodyne method,* and an arrangement of this kind is described in abstract No. 18.2. In this latter case, however, a number of further possibilities are available. There are, for example, a number of audio-frequency meter-circuits in which the response (direct current, or deflection of a moving element, as the case may be) depends only on the audio-frequency of the input voltage and not on its amplitude. Such an arrangement is described by F. GUARNASCHELLI and F. VECCHIACHI in the *Proceedings of the Institute of Radio Engineers* (Vol. 19, pp. 659–663), and another is described by E. and R. MITTELMANN and M. WALD in the *Zeit. für. Hoch.* (Vol. 37, pp. 187–199). The applicability of

* Since the above was written a frequency-controlled relay of this kind, capable of varying the tuning condenser of the transmitter, has been designed and constructed by H. A. Thomas at the National Physical Laboratory. It is described in Provisional Patent Specification No. 2446/33.

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such arrangements to automatic monitoring is worthy of investigation.

In this connection it should be noted that there would probably be considerable advantages in utilising a fairly high harmonic of the transmitter to give the heterodyne beat with the reference-oscillator. Not only would the sensitivity of control be enhanced thereby, but the tendency of the transmitter to pull the frequency of the local oscillator would be eliminated. This would also facilitate the control of long-wave transmitters by reference to a short-wave crystal-controlled local-oscillator.

Some of the automatically-monitored schemes described in current literature (abstracts 17 and 18.1) do not embody any provision for making the controlling force independent of oscillation-amplitude, and this provision may not, in fact, be a necessity. It would seem, however, to be a very desirable feature if it can be obtained without undue complication of equipment.

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PART II

Annotated Abstracts of Selected Published Papers and Official Memoranda on Valve-Oscillators of Stable Frequency

1. **Constant-Frequency Oscillators.** F. B. Llewellyn: *P.I.R.E.*, 1931, **19**, 2063-2094

Abstract

The paper is concerned with circuit-arrangements designed so that none of the physical constants of the valve, other than the interelectrode capacitances, appear in the calculated value of the frequency of oscillation, which is therefore that of the oscillatory circuit alone, as modified by the interelectrode capacitances.

It is assumed throughout that the oscillatory-circuit resistances are negligible compared with those introduced into the circuit by connection to the valve.

It is shown analytically that for all the usual forms of oscillating circuit, the insertion of suitable impedances into the grid and anode connections will give oscillation-frequencies which are independent of all valve constants other than interelectrode capacitance. It is also shown that similar results can be obtained by circuit-arrangements which are, in effect, such as to give a coupling-coefficient of unity between the grid and anode circuits.

Experimental results are given which show that, in fact, circuits designed on the above lines give an exceedingly high degree of frequency-constancy, of the order of a few parts in a million in some cases, with respect to deliberately imposed variations of supply-voltages.

Comments

The work is of real value in showing how one source of variation (*i.e.*, changes in valve-constants due to ageing or variation of supply-voltages) can at least be minimised in certain types of master-oscillator. It should be noted, however, that the experimental results were obtained with valves of very small power-output.

The principal application would seem to be to master-oscillator systems operating at one fixed frequency. In nearly all cases the circuits are not very conveniently adapted to covering a range of frequencies in a band or to appreciable changes of frequency in steps.

The interelectrode capacitances are assumed to be constant. In actual practice they may be expected to vary somewhat with conditions of oscillation, but it appears that such variation would

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not invalidate the compensation methods described, though it would affect the actual values of the appropriate correcting impedances.

2.1. **Description of a Heterodyne Wavemeter.** E. FROMY: *L'Onde Elect.*, 1925, **4**, 433-441

2.2. **Frequency-Stabilisation of Valve Oscillators.** E. MALLETT: *J.I.E.E.*, 1930, **68**, 578-582

2.3. **Frequency-Variations in Thermionic Generators.** K. E. EDGEWORTH: *J.I.E.E.*, 1926, **64**, 349-360

Abstract

The papers deal with small-power oscillating circuits (master-oscillators or heterodyne wavemeters) in which a special disposition has been adopted to ensure that the valve constants do not affect the frequency of sustained oscillation.

The paper by Edgeworth gives, in addition, useful information about the nature of the observed frequency-variations and an instructive analogy based on forced oscillation in a pendulum, pointing out the importance of the phasing of the power-supplying impulses.

In the Fromy circuit a high degree of stability with respect to supply-voltages is attained (parts in 100,000) by an arrangement involving three tuned circuits.

Comments

The above papers describe particular circuit-arrangements which satisfy the general conditions given in the paper by Llewellyn (abstract No. 1), and the same general comments apply. A very high degree of stability is claimed for the Fromy circuit, but it is somewhat complicated in form, and simpler means are available for producing comparable results.

2.4. **On Some Processes for the Stabilisation of the Frequency of Radio-Oscillators.** MARC BRUZAU: *L'Onde Elect.*, **11** (1932), 297-314

Abstract

The paper refers to oscillation maintained by means of a negative-resistance effect obtained in a double-grid valve (not a screen-grid valve). The negative resistance occurs between the filament and a common connection to the two grids, and may be made as low as about 4,000 ohms. It is pointed out that this negative slope arises as a consequence of variation of distribution of primary electron emissions between the various electrodes, and does not depend on secondary emission.

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The means of stabilisation examined analytically and experimentally are :—

(*a*) insertion of inductance between the oscillatory circuit and the negative resistance ;
(*b*) equalisation of the resistance in the inductive and capacitive branches of the oscillating circuit ;
(*c*) division of the inductance into two equal parts, with zero mutual inductance, and the connection of the negative-resistance terminals to the centre point and one end of the inductance.

Experimental examination of these arrangements was carried out with a Philips double-grid valve A441. The circuit used was that known as the Neuman circuit, *i.e.*, outer grid connected to the inner grid by a condenser (100 $\mu\mu$F), inner grid through the oscillating circuit to + 60 volts, anode to + 30 volts. The frequency was about 850 kilocycles per second. For the first method of stabilisation it was found that an initial frequency-variation of about 0·1 per cent. for an 8 per cent. variation of filament-current was reducible to about 0·01 per cent. by inserting an inductance equal to a quarter of that of the oscillating circuit in the connection to the negative resistance. The theoretical compensating inductance should be equal to that of the oscillating circuit. The difference is attributed to the effect of resistance in the condenser branch. Similar results, *i.e.*, stability to about 0·01 per cent. for 6–8 per cent. changes of filament-current, were obtained by the other means described above. In the third method, the stabilised condition was with 35 turns in the inductive branch and 19 in the condenser branch instead of with equality of these inductances.

Comments

The theory underlying the above processes of stabilisation is given in section 2.3.1 of the essay.

The agreement between theory and experiment is qualitative only, the quantitative discrepancy being considerable. The smaller than theoretical value of the stabilising inductance is attributed to resistance in the capacitive branch, but the analysis shows that the resistance of the condenser would have to be 3/5 of that of the inductance to account for the difference. It is suggested that the discrepancy may be due to a variation of interelectrode capacitance with supply-voltage and with the oscillating condition, a factor not included in the analysis. The quantitative discrepancy does not, of course, detract from the utility of the stabilisation-methods described.

It is stated that secondary emission plays no part in the negative-resistance phenomena utilised in this circuit. Further experimental evidence on this point would be desirable, as the electrode-voltages

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are such as to make possible secondary emission from the anode to the grids.

No data are given as to frequency-stability with time, supply-voltages being kept constant.

The circuits and methods described appear to be applicable to small-power master-oscillators only.

3. **On the Stabilisation of the Frequency of a Heterodyne Oscillator over Wide and Continuously Variable Ranges of Frequency.** N. K. TITOV and A. I. VAINBERG: *J. App. Phys.* (*Mos.*), **7**, 37-60. Also (in French) in *Documents du C.C.I.R.* for 2nd Meeting, 1931, 129-146

Abstract

The paper contains a very detailed analytical examination of the self-maintained oscillator, the dependence of frequency on each element of the assembly being studied separately and means for its minimisation being deduced. As applied more particularly to the Meissner circuit (tuned anode) the following are concluded to be conducive to frequency-stability :—

(*a*) small L/C ratio ;
(*b*) valve of high slope resistance ;
(*c*) high non-inductive resistance (100,000 ohms or so) in the anode-circuit ;
(*d*) condenser and grid-leak in grid-circuit.

It is pointed out that the chief function of the high resistance in the anode-circuit is the maintenance of appropriate phase-relationships between the grid- and anode-voltages, and that, though theoretically less perfect than the series impedance arrangements elsewhere described, it has the advantage of being substantially independent of frequency.

As far as the back-coupling conditions are concerned, small grid-inductance and loose coupling are advocated. (See comments on this point.)

A considerable amount of experimental work is described, supporting the above conclusions, both at audio and radio frequencies (up to about one megacycle), using small receiving valves. A circuit embodying the features listed above is claimed to show stability of the order of 0·001 per cent. for wide ranges of variation of supply-voltage.

Comments

The paper is most valuable as a very detailed and thorough analytical and experimental investigation of the small-power master-oscillator. The experimental data in particular are of great value in showing how very important the individual factors and their proper adjustment may be.

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The conclusions with regard to coupling may perhaps be criticised as being not very clearly and definitely established. Other evidence suggests that the desirable condition is not loose coupling between the grid- and anode-circuits, but small mutual with close coupling, *i.e.*, the minimum grid-inductance coupled as closely as possible. (*Cf.* paper by Llewellyn.) This point appears to call for further experimental investigation.

4. **On the Instability of the Frequency of Valve Generators, and their Stabilisation.** W. LAZAREF: *Zeit. für. Hoch.*, 1929, **33**, 55-63

Abstract

The nature of the frequency-variations of a simple back-coupled tuned anode-circuit oscillator with coupling and with supply-voltages are examined theoretically and experimentally, special attention being given to the effect of grid-current. The latter is shown to be a very important factor. It is stated that the simplest and most effective means of stabilisation is the elimination of grid-current. A circuit so stabilised is described. It consists of a double-grid valve of which the outer or " control-grid " is left free, and the inner, described as the " space-charge " grid, is connected to the back-coupling inductance through a small capacity of high insulation, but has no d.c. connection to the filament. At frequencies of about 500 cycles per second, and up to 3×10^6 cycles per second, it is stated that the stability of frequency with respect to filament- and anode-voltages is of the order 0·003 per cent. and 0·0001 per cent. respectively. It is stated that the coupling should be loose, to avoid " irregularities " of oscillation.

Comments

The analysis and experimental work do not bring out any points not found in the other papers listed and the abstract is included mainly as an illustration of the degree of stability obtainable by very simple means. It appears that only small-power master-oscillator or heterodyne circuits are contemplated. It is not clear why the particular type of valve was used, nor whether an ordinary triode would give similar results. The " irregularities " of oscillation referred to are presumably due to a state of intermittent oscillation, sometimes referred to as " squegging."

5.1. **Vacuum-Tube Oscillators.** J. W. HORTON: *Bell Syst. Tech. J.*, 1924, **3**, 508-524

Abstract

Systems for the maintenance of continuous oscillation by regeneration are considered in general form. It is shown that all

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such systems are reducible ideally to two main elements—a frequency-control unit (*e.g.*, tuned circuit or other form of resonator) and an amplifier unit, in general not critical with respect to frequency.

The power, amplitude and phase relationships appropriate to the maintenance of stable oscillations by such systems are then discussed in general terms, and illustrated by reference to a two-valve system for the generation of oscillations. The conclusions arrived at are applied to the design of a very simple form of oscillator suitable for frequencies from 100 to 100,000 cycles per second, consisting of a shunt-fed tuned anode-circuit with a series resistance in the connection to the anode, and a closely-coupled grid-winding. Frequency-stability of the order of 0·03 per cent. with respect to supply-voltage changes and load changes is claimed.

Comments

The paper is chiefly of value as giving a very clear and generalised enunciation of the fundamental principles of oscillation-generators, in graphical and physical terms. It is exceptionally lucid and useful in this respect. The most obvious application is to master-oscillators of small power-output. In point of theory the paper is similar to that by Llewellyn (abstract No. 1), giving in physical terms the ideas which are developed somewhat more completely and generally in analytical form by Llewellyn. It is interesting to note that even with the very simple circuit described frequency-stabilities of the order of parts in 10,000 are obtained in small-power circuits, made practically independent of load by a form of resistive coupling to the latter.

5.2. Quantitative Analysis of the Oscillations in Valve Generators. H. G. MÖLLER: *Jahrbuch. Tel.* 1919, **14**, 326-360

Abstract

The method of analysis is purely graphical, and the physics of the problem is stated in language practically identical with that used by Horton (abstract No. 5.1). The method depends on the construction of " oscillation-characteristics " for the valve, giving the vector relation between the fundamental component of the anode-current and the anode-voltage for assigned values of anode-voltage and grid-voltage variation. The families of characteristics so obtained are then related to similar characteristics applying to the tuned circuit, and the stable state of oscillation is deduced as an amplitude and phase balance between these two sets of characteristics.

Further, an interesting stable-frequency circuit is mentioned. The L and C branches are made of equal resistance, and the grid-winding is coupled to each branch equally.

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Comments

The paper is of interest historically, as being apparently the first attempt at a rigorous solution of the oscillating state in terms of actual as distinct from ideal rectilinear characteristics. From a practical point of view, however, the method is exceedingly laborious and is, moreover, of doubtful validity at very high frequencies, where, on account of displacement currents through the inter-electrode capacitances, the static characteristics no longer suffice as data for determining the total flow of current through the valve in terms of assigned anode and grid potentials.

The principal value of the method would seem to be in giving a qualitative interpretation of observed behaviour, a comment equally applicable to the similar mode of description given by Horton (abstract No. 5.1).

6. **Frequency Variations in Valve Oscillators.** D. F. MARTYN: *Wireless Engineer*, 1930, **7**, 3-15

Abstract

Experimental data are given of the frequency-variations of a tuned anode-circuit oscillator, with coupling, circuit resistance, and with supply-voltages. A theory is given which takes the effect of grid-current into account, and it is stated that all the observed variations can be attributed to the effect of grid-current. It is also pointed out, however, that interelectrode capacitance, being dependent on space-charge (and also on temperature) may also vary with operating conditions and thus cause small frequency-variations, particularly at the lower frequencies.

For the minimisation of frequency-variations it is suggested that (i) the anode-voltage and negative grid-bias should be as large as possible, and the filament-voltage as low as possible; (ii) the coupling close and the grid-winding small in inductance. Such an oscillator is claimed to remain steady to one in a million over a period of hours, but no details are given as to variation with supply-voltages or other factors. A constancy of a few parts in 10,000 is given as a possible stability with respect to supply-voltages on the basis of results published by other writers.

Comments

The paper is useful in calling attention to the importance of suppressing grid-current in respect of frequency-stability, but seems to under-estimate the importance of other factors, such as suitable phase relationships between grid and anode voltages. It is also useful in calling attention to the part played by variation of inter-electrode capacitance with conditions of operation. The statement that the importance of this latter variation decreases with frequency and is " entirely negligible " at " high-frequencies " is one on which further experimental data are desirable.

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7. A New Frequency Stabilised Oscillator System. R. GUNN: *P.I.R.E.*, 1930, **18**, 1561-1574

Abstract

The paper describes a master-oscillator circuit. It is essentially similar to the standard multivibrator circuit, but with syntonised tuned circuits in the anodes instead of the resistances of the multivibrator, and with very complete electric screening between the two valves and their associated components. This is obtained partly by ordinary metallic shielding and partly by the use of screen grid valves. For this arrangement frequency-stability of the order of parts in 100,000 is claimed for 10 per cent. changes in the supply-voltages.

Various modifications of the circuits are described. Reference is also made to the use of a small auxiliary condenser with one bimetallic plate as a means of compensating for changes in circum-ambient temperature.

Comments

Although the circuits are described as very flexible, and preferable to quartz circuits in this respect, the description suggests that very great care is needed in their design and manipulation, particularly in the matter of coupling to the amplifier. It is emphasised that a very restricted power-output is essential for very high stability.

The theoretical explanation is couched in terms of "circulating" and "re-entrant" oscillations, and is not very easily understood. The system may be considered as a whole to be in a state of stationary oscillation, the circuit conditions being such that the power-supply to each tuned circuit is correctly phased. In fact, one of the circuits described, which is a single valve tuned-grid tuned-anode arrangement, with a tuned circuit coupling between them, makes it probable that this simpler explanation is adequate.

It may be doubted whether the systems described are the simplest means of obtaining the recorded master-oscillator stabilities. On the other hand, it is claimed that the circuits have found wide application in naval and commercial aircraft.

8. A Recent Development in Vacuum Tube Oscillator Circuits. J. B. DOW: *P.I.R.E.*, 1931, **19**, 2095-2108

Abstract

The paper describes a master-oscillator output circuit combination, using a four-electrode valve, for which a very high degree of frequency-stability is claimed with respect to supply-voltages and load-variation.

The oscillatory circuit itself is essentially of the Colpitts type, the mid-condenser point being earthed, and the outer terminals con-

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nected to the control-grid and screen-grid respectively of a screen-grid valve. The more distinctive feature of the arrangement is that the output potential variation is derived from a resistance or impedance connected in the normal or outer anode-circuit. This is described as an " electron-coupling " between the two systems.

As far as the master-oscillator part is concerned, the very high frequency-stability with respect to supply-voltages claimed for this (parts in a million) appears to be due to a mutual compensation of effects of changes of the voltages of the screen-grid and outer anode. A common voltage-supply for these electrodes is a condition of this stability.

A very high degree of freedom from frequency-variations due to change of load-conditions is claimed, and attributed to the lack of direct connection between the output circuit and the oscillator.

Temperature-variations are about 0·005 per cent. per degree Centigrade without any temperature-control and about 0·001 per cent. per degree C. with temperature-control of the master-oscillator unit.

It is claimed that the overall performance of transmissions controlled by a master-oscillator of the above type has been reported as " pure direct-current crystal-control " at frequencies up to 24,000 kc/s.

Comments

The so-called " electron-coupling " certainly appears to be a very convenient means of transfer of potential to an output-circuit, though in one of the forms described accurate neutralisation of the screen-grid to anode capacitance is a condition of success. This balance is not likely to be quite independent of frequency. An alternative arrangement minimises this undesired coupling-capacitance by using a form of pentode valve.

Apart from any undesired coupling by capacitance, the coupling mechanism can properly be described as uni-directional, inasmuch as changes of screen-grid and control-grid voltage produce changes of voltage in the anode circuit load impedance, whereas changes in this output voltage will not, to a first order, produce any corresponding changes in the input voltages.

Ideally speaking, the same should be true of a suitably designed and neutrodyned coupling through the agency of a separate coupling valve, but the Dow arrangement may be a particularly convenient disposition for the purpose.

9.1. A Master Oscillator of High Constancy of Frequency.
T. D. PARKIN : *Marconi Review*, July, 1930, 14-21

Abstract

The article describes a temperature-controlled small-power master-oscillator for operation at frequencies of the order 10^3 cycles per second. The oscillator circuit itself consists of an inductance

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and variable condenser, the terminals of which are connected to the grid and anode of the valve through condensers of small capacitance. There is no centre tap, and the H.T. supply is shunt-fed through a resistance. The oscillator is mounted inside a lagged wooden case lined outside with copper. A very simple form of thermostat regulator is used. It consists of a steel plate carrying steel blocks to which are secured the ends of a bowed strip of some other metal of different expansion. Contact is made on the centre of this strip and controls heating current through a relay. The condenser is of air dielectric series type, and the inductance of copper, wound on porcelain.

No details are given of the method of coupling to the transmitter, but it is stated that one frequency doubling stage is employed.

The maximum frequency-variation is stated to be about 2 in 100,000 for 5 per cent. changes of supply-voltages, and, after an initial drift lasting 4 hours, the constancy in a 6-hour run was $\pm$ 10 in a million.

Comments

The paper is included as an indication of performance attainable with comparatively simple apparatus. The stability of the circuit itself appears to depend on the elimination of grid-current, loose capacity-coupling to the valve and probably (though details are not given) on careful choice of L and C. It does not appear to embody any special device. The components are not themselves temperature compensated, but the temperature of the whole outfit is stabilised by a thermostat. It appears that such an oscillator would need to be run continuously as there is appreciable initial drift lasting a few hours.

9.2. **Frequency Stabilisation of Triode Oscillators.** *Post Office Memorandum Radio D.H.* 336*

Abstract

A high degree of constancy of the generated frequency of a triode-oscillator with variation of the supply-voltages may be achieved as follows :—

(1) By minimisation of the amount of grid-current in the oscillator-valve by the use of a grid-leak and condenser; this is a convenient alternative to the complete elimination of grid-current, which in general necessitates a very high d.c. anode-potential;

(2) By adjustment of the capacitance of the grid-condenser, inductance of grid coil and mutual inductance of grid and tuned anode coil, and by using a suitable series anode-circuit resistance;

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(3) By employing a circuit of low decrement with low L/C ratio, the circuit-elements being of robust and stable construction.

The above methods have only been applied to oscillators functioning over a small range of frequencies of the order 55–65 kc/s. There would appear to be no reason why the methods should not be applied at other frequencies.

Results obtained are as follows :—

Valve V.T.65. Shunt-fed tuned anode-circuit, with series "feed-back" resistance

Filament current change	*Frequency change*
− 0·07 amps., *i.e.*, − 9 per cent.	− 0·002 per cent.
+ 0·05 amps., *i.e.*, − 6 ,,	− 0·002 ,,

Anode supply-voltage change	*Frequency change*
− 10 volts, *i.e.*, − 8 per cent.	+ 0·0008 per cent.
− 5 volts, *i.e.*, − 4 ,,	+ 0·0007 ,,
+ 5 volts, *i.e.*, + 4 ,,	− 0·0005 ,,

Over a run of 18 hours, during which the room temperature changed 4°, the drift in frequency was only about 1 cycle in 60,000. It appears that some fortuitous temperature compensation is operative in the inductance and capacitance.

Comment

The above is an interesting example of an exceedingly high degree of frequency-stability in a small-power oscillator, obtained partly by care in initial design and partly by adjustment, apparently on an empirical basis, of circuit elements. The chief stabilising factors appear to be the grid-capacitance, grid-leak, and series feed resistance. (*Cf.* Part I, section 2.3.2, paragraphs i, ii and v, p. 6.)

9.3. **The Frequency-Constancy of the Beating Oscillator used in Post Office Short-Wave Radio Receivers.** *Post Office Memorandum Radio D.H. No.* 354*

Abstract

The circuit is a two-valve tuned-anode push-pull arrangement with cross-couplings by capacitance from anodes to grids, and with a series resistance included in the H.T. supply-lead to the coil-centre. Frequency range 8–23 megacycles per second.

The circuit-elements (not the valves) are housed in a celotex-lined metal chamber to produce some degree of temperature-stabilisation. The observed frequency-variations for 5, 10, 15, 20, 25 per cent. changes of filament-voltage, and 5, 10, 15, 20, 45 per cent. changes of anode-voltage are shown as a series of curves plotted against operation-frequency over the range 8–23 megacycles per second. Taking,

* See footnote on p. 40.

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for example, the curve for 10 per cent. decrease in filament-voltage, from 9–12 megacycles per second, it rises uniformly to 12 parts in 10^6, and from 12–18 megacycles per second falls nearly uniformly to — 134 parts in 10^6, passing through zero at 13 megacycles per second. The lines for 5 and 15 per cent. decreases of filament-voltage are similar in shape and also pass through zero at 13 megacycles per second. At this frequency, therefore, there is no frequency-change with filament-voltage. For 10 per cent. decrease in anode-voltage the shape is very similar, a rise to 40 in 10^6 at 12 megacycles per second falling at a decreasing rate to — 20 in 10^6 at 18 megacycles per second. The lines for 5 and 15 per cent. decreases are similar in shape, but they do not intersect at zero change as in the filament-voltage case. There is not, therefore, any point of complete stability with respect to both supply-voltages.

With the anode series resistance short-circuited, the filament-voltage frequency-variations are increased about three-fold, but the anode-voltage frequency-variations are not greatly affected.

Comments

There seems to be no special stabilising element in the above design, though grid-current effects are minimised by the capacitance coupling of the valves, which has an effect equivalent to grid-leak and condenser action. The particular value of the memorandum lies in the complete information given over the whole operating-range of frequencies. The effect of any given supply-voltage variation is shown to depend to a great extent and in a complicated manner on the frequency of operation, *i.e.*, on the tuning-capacitance for the given coil, and it is clear that where variable-range small-power oscillators are concerned information relating to a particular frequency will only be of correspondingly limited significance and may even be quite misleading. Thus observations at 13 megacycles per second would have suggested that the oscillator frequency was quite independent of filament-voltage, whereas at 18 megacycles per second there is a change of 90 parts in a million for 10 per cent. filament-voltage change.

10.1. **Chairman's Address, Wireless Section, I.E.E.** C. E. RICKARD: *J.I.E.E.*, 1930, **69**, 11-24

Abstract

The address contains a section headed "Frequency-Control," and a sub-section on "Valve Master-Oscillators." The principal difficulties in this type of control are ascribed to changes in the oscillatory-circuit components due to temperature, and also inconstancy due to "flow of the material." It is stated that "several circuits are known which take care of the valve problem." On the other hand, reference is made to the lower standard of stability

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given under service as compared with laboratory conditions. Data are given of initial frequency-drift with thermo-stated master-oscillator controlled systems of normal type (*i.e.*, previous to the introduction of the Franklin drive). In the first 20 to 30 minutes this may amount to a fall of 0·05 per cent., and appears to be approximately independent of frequency.

The Franklin drive is referred to briefly. In this system the oscillatory circuit is of robust construction, and a self-compensation of the circuit as a whole for temperature-changes is obtained by using differential expansion between the coil-former (a loaded ebonite material) and a brass rod to produce a suitable correcting capacitance-change. This is claimed to give constancy to 1 in 10^6 per degree, and is used without temperature-control. The anode-voltage effect is given as about 1 in 100,000 for a $2\frac{1}{2}$ per cent. change, and the filament-voltage effect as about the same for a 10 per cent. change. The frequencies concerned are of the order 10^7 cycles per second.

Comments

It appears that a high degree of constancy has been obtained with the Franklin type of drive. The actual drive-circuit is not described in the address, but a description is given in the publication referred to in abstract 10.3.

The circuit temperature-compensation is as a unit, *i.e.*, the coil and condensers are not separately compensated. It is not stated whether this compensation is appreciably affected by frequency-adjustment.

10.2. **Chairman's Address, Wireless Section, I.E.E.** A. S. ANGWIN : *J.I.E.E.* (1931-32), **70**, 17-35

Abstract

The address is mainly concerned with means of frequency-measurement and control, but is included among these abstracts because it contains observed frequency-variation figures for certain actual master-oscillator controlled transmitters.

For two master-oscillator controlled, high-speed, short-wave telegraph transmitters on allocated frequencies of 18,100 and 9,260 kilocycles per second the observed frequency-variations over a period of about six months were 7 and 15 kilocycles per second, *i.e.*, 0·04 per cent. and 0·15 per cent. respectively. A medium-wave master-oscillator controlled broadcasting transmitter on 1,148 kilocycles per second showed variations not exceeding 50 cycles in a day's operation, *i.e.*, 0·005 per cent. A low-power shore-station with a simple self-oscillating circuit, nominal frequency 460 kilocycles per second, showed variations of less than 0·1 per cent. in a period of about 8 hours. For a short-wave ship transmitter with master-oscillator control on a nominal frequency of 12,380 kilocycles

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per second, the observed variation over a period of about four months was just less than 0·04 per cent. In another case on the same nominal frequency the observed variation over about four days was ± 1·1 kilocycles per second, *i.e.*, 0·009 per cent.

10.3. **Short-Wave Wireless Communication.** A. W. LADNER and C. R. STONER: Text Book: Chapman & Hall, 1932

Abstract

The above recently-published book contains a chapter on constant-frequency oscillators (Chap. IX, pp. 150–173). In particular, a brief account is given of the Franklin master-oscillator. The oscillating-circuit part of this has already been mentioned (abstract 10·1). It is driven by two triode valves. The oscillating circuit is coupled to the grid of the first valve by a very small capacitance (about 1 $\mu\mu$F). This valve is coupled by resistance and capacitance to the second, described as a " phase-reversing valve," having a pure-resistance anode-load. The anode of the second valve is coupled to the free end of the oscillating circuit, and thus to the grid of the first valve, by a very small capacitance (about 1 $\mu\mu$F).

The advantages of the circuit are stated to be (*a*) very loose coupling between oscillating circuit and valves, made possible by the amplification given by the second valve; (*b*) the possibility of earthing one end of the oscillating circuit.

These driving circuits are usually made for wave-lengths between 100 and 200 metres, and a harmonic component is selected for the control of the transmitter. It is claimed that the " precision " of this drive is " equal to that of a good crystal " and that it is " self-compensating as regards temperature over a range of 40° C."

Comments

Particulars of frequency-stability for this type of circuit are given in abstract No. 10.1. The circuit is further discussed in section 2.4 of Part I (p.).

11.1. **The Form of Free Triode Oscillations.** E. V. APPLETON and B. VAN DER POL: *Phil. Mag.*, 1921, **42**, 201-219

11.2. **Oscillation-Hysteresis in Simple Triode Generators.** E. V. APPLETON and B. VAN DER POL: *Phil. Mag.*, 1922, **43**, 177-193

11.3. **The Triode as a Generator of Oscillations.** B. VAN DER POL: *Rad. Rev.*, 1920, **1**, 701-710 and 754-762

Comments

The above three papers are similar in kind to those mentioned in abstracts Nos. 5.1 and 5.2, *i.e.*, they deal with the analysis of oscillatory systems, valve-maintained, without restriction to linear

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operation. Their principal importance in the present connection lies in their use of an " oscillation-characteristic," which is a relation between anode-current and anode-voltage, having due regard to the relation between anode-voltage and grid-voltage. That is, it is the eliminant of v_g from the equations

$$i_a = f(v_a + \mu v_g)$$

and

$$v_g = \chi(v_a).$$

The chief utility of this mode of analysis is that it makes the triode oscillator analytically similar to the simpler negative-resistance or dynatron type of oscillator.

It may be doubted whether the analyses given are applicable as they stand to, for example, triode-maintained oscillation at high radio frequencies, but they contain fundamental ideas of considerable value in the theory of oscillation in general.

12.1. **Operating Mechanisms of Negative-Resistance Oscillators.** R. USUI: *J.I.E.E. Japan*, 1931, **51**, 64-67 and 115-116

12.2. **Analytical Studies on the Fundamental Characteristics of Triode Oscillators.** R. USUI: *J.I.E.E. Japan*, 1932, **52**, 44-47

12.3. **Study of Sustained Oscillations.** A. LIENARD: *Rev. Gen. d'Elec.*, 1928, **23**, 901-912 and 946-954

12.4. **Valve Oscillators Adjusted to Near the Oscillation Threshold.** Y. ROCARD: *Comptes Rendus*, 1932, **194**, 1325-1327

12.5. **Study of Weak Oscillations of Valve Generators.** M. BIASI GIOVANNI: *L'Onde Elec.*, 1932, **11**, 89-100

12.6. **On the Solution of the Representative Differential Equation of the Triode Oscillator.** E. V. APPLETON and W. M. H. GREAVES: *Phil. Mag.*, 1923, **65**, 402-414.

Comments

The above papers have been listed together because they are all concerned with the analysis of oscillatory systems in which stable amplitude of oscillation results from the non-linearity of the system. That is to say, they represent an attempt at the analysis of the behaviour of actual systems as distinct from the simplified and idealised linear systems usually considered.

The papers are chiefly concerned with the general equation

$$\ddot{x} + f(x)\,\dot{x} + \omega_0^2 x = 0,$$

which is fundamental in the theory of self-maintained oscillations of finite amplitude. They contain matter of fundamental importance

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to the study of actual as distinct from ideal oscillatory systems, but cannot usefully be abstracted. They are referred to more fully in section 3.1 of Part I.

13. **The Dynatron Oscillator.** J. GROSZKOWSKI: *Wireless Engineer*, 1932, **9**, 446-447

Abstract

The above is a letter, preliminary to a further publication, which has not yet appeared.* The writer points out the inadequacy of the simplified analyses of oscillatory systems confined to operation on linear characteristics. He intimates that an analysis based on energy-changes is required for the more general and more practical cases. In particular, he states a result relating frequency with total harmonic content derived, apparently, from the condition of balance of "imaginary energy" (*i.e.*, presumably the so-called "watt-less power") within the tuned circuit.

Comments

The above contribution, though brief, may prove to be of importance, particularly from a theoretical point of view, as calling attention to a somewhat neglected analytical method of study of oscillatory circuits. It is not improbable that analysis based mainly on energy-considerations may prove more readily applicable to actual valve-conditions than the usual amplitude-analysis.

The general principle involved is discussed more fully in sections 3.2 and 3.3 of Part I.

14. **On the Influence of Valve Resistance in Oscillation Generators.** N. W. MCLACHLAN: *Wireless Engineer*, 1932, **9**, 130-135

Abstract

The main object of the paper is to point out that analysis of oscillatory systems based on the limiting condition for the incidence of oscillation will tend to misleading conclusions as to frequency-variations and the means of minimising them, since the actual state of stable oscillation will necessarily produce changes in the system which invalidate the analytical assumptions. In particular, the circuit-arrangement proposed by Mallett (abstract No. 32.2) is criticised from this point of view.

Another section of the paper of considerable practical interest describes means of varying the effective negative resistance of

* Since the above was written, the paper by Groszkowski has been published in the *P.I.R.E.* for July, 1933 (Vol. 21, pp. 958–981) under the title "The Interdependence of Frequency-Variation and Harmonic Content and the Problem of Constant Frequency Oscillation."

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dynatron types of oscillatory circuit. In particular, it is pointed out that the inclusion of a positive resistance in series with the anode or negative resistance circuit will have this effect.

Comments

While the main contention of the paper must be admitted, it may nevertheless be pointed out that investigation of the condition of incidence of oscillation, and the application of methods of frequency-stabilisation based on this analysis, have justified themselves experimentally, as illustrated, for example, by the circuit-arrangements discussed by Llewellyn (abstract No. 1).

The ideas contained in the constructive sections of the paper, particularly that of reducing the magnitude of the negative resistance of a dynatron circuit by the inclusion of a suitable positive resistance, have practical implications which are by no means exhausted in the paper.

15. **Low-Frequency Oscillator Stabilised by a Tuning-Fork.** G. MABBOUX: *Comptes Rendus*, 1931, **192**, 1154-1156

Abstract

A wound bobbin, mounted between the prongs of a tuning-fork, is included in the grid-lead of a series-fed Hartley circuit, tuned to the same frequency region as the tuning-fork. Over an appreciable range of adjustment of the capacity of the oscillatory circuit the frequency is controlled by the fork. It is claimed that a high degree of stability is obtained with respect to variation of supply-voltages.

Comments

This item appears to have no very direct bearing on the stable self-oscillation of radio-frequency circuits, but is included because, in common with a similar quartz-crystal circuit referred to in abstract No. 17, it suggests the possibility of frequency-control by means of an associated circuit which is very lightly loaded and loosely coupled to the source of energy as compared with the circuit in which the bulk of the oscillating energy is dissipated.

16. **New Methods of Frequency-Control Employing Long Lines.** J. W. CONKLIN, J. L. FINCH and C. W. HANSELL: *P.I.R.E.*, 1931, 1919-1930

Abstract

The paper describes oscillator-circuits in which transmission-lines of low decrement are utilised in a similar manner to quartz-crystals for the stabilisation of frequency, the principal advantage

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to be gained thereby being the large " volt-ampere " capacity of the transmission-lines as compared with quartz-crystals and the consequent elimination or reduction of the amplification required for the operation of a transmitter.

The circuits are described under two main headings. The first covers a number of arrangements which differ only from already known quartz-controlled circuits in the substitution of a loop transmission-line for the crystal. As a practical example, a description is given of a transmitter (WIK) at 13,930 kc/s. (two 20 kw. water-cooled valves in push-pull, directly coupled to the antenna). Improved reliability is claimed for this as compared with a previous crystal-control. In another case (35,000 kc/s.) frequency-constancy was within $\pm$ 0·02 per cent. for long runs. The observed variations were mainly due to temperature-change (about 10° C.).

Under the second heading are described circuits which are considered to operate on somewhat different principles. They involve tuned grid- and anode-circuits which are normally non-oscillatory (due to screening or neutrodyning), but which are made to oscillate by coupling through tuned transmission-systems. This is called " aperiodic " long-line frequency-control, which presumably may be taken to mean that the line is not considered to be in a state of free oscillation, as it is in the previous cases. It is stated that the circuits have been applied to actual transmitters, but no details of performance are given.

Comments

The paper gives very little detailed information and no detailed analysis. The high stability recorded may be partly due to the low decrement of the long line considered as an oscillating circuit, but is probably also due to the special nature of the oscillatory circuit in respect to harmonics, *i.e.*, the circuit has zero reactance not only to the fundamental frequency, but also (very approximately) to all harmonic frequencies. For the reasons given in section 3 of Part I, this should minimise the effect of wave-form on frequency. Another factor of importance is that the interelectrode capacitance should have comparatively little effect on frequency in this type of circuit.

From the account given it seems that the fundamental idea is valuable and is likely to find useful application. It is clear, however, that its range of application is likely to be limited to comparatively short wave-lengths owing to the lengths of transmission-line involved. It does not at present appear possible to obtain any appreciable compression of these lines without loss of frequency-stabilisation, but this possibility should be further explored.

The scheme should have a useful degree of elasticity in frequency-adjustment, and there appears to be no inherent power-limitation.

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17. Frequency-Stabilisation of Radio Transmitters. Y. KUSONOSE and S. ISHIKAWA: *P.I.R.E.*, 1932, **20**, 310-339

Abstract

The paper is divided into three main parts:

The first deals with two special types of circuit intended for the minimisation of frequency-changes due to the valve itself. In one of these a variable resistance (of the order 10^2–10^3 ohms) is common to the grid- and anode-circuits, arranged so that an increase in mean anode-current will tend to lower the mean grid-potential. A fairly high value of grid-leak resistance is used in addition. Experimental results obtained are: with a receiving valve at 100 kc/s., frequency-constancy to 0·001 per cent. for wide variations of supply-voltage; with a small power-valve (250 watts) constancy to about 0·005 per cent., radio-frequency not stated. In the power valve experiments a steady frequency-drift up to about 0·05 per cent. was observed, lasting through the first hour of operation. This is attributed partly to change of circuit-constants by rise of temperature, but mainly to a change of valve interelectrode capacitance.

The other of the two types of circuit dealt with in the first part of the paper is based on the fact that frequency-variations in a Hartley circuit are of opposite sense to those arising from the same cause in a similar Colpitts circuit. The circuits described make use of centre-tapped inductances and centre-tapped condensers. The centres are connected by a resistance (or, in some cases, an inductance) and earth connection is made to a tapping on this centre-connecting element. The arrangement is described as "phase-compensation" on the grounds that its action is to maintain suitable phase-relationships between the grid- and anode-voltages. Experimental results are consistent with the theory and show frequency-constancy of the order of 0·005 per cent. with a small receiving valve for a wide range of variation of supply-voltages, and of about 0·02 per cent. for a 100 per cent. change in anode-voltage in the case of a 1·5 kw. valve at about one megacycle per second.

The second part of the paper deals with what appears to be a new form of crystal-controlled oscillator, the object of the circuit being to obtain an appreciable increase in the permissible output from the controlled circuit. A parallel-tuned circuit is connected in series with the grid-lead of an oscillator of ordinary type, the connections being to one end of, and to a tapping on, the inductance of this inserted circuit. The crystal is connected across the inserted circuit. An experimental case described in detail is that of a 100-watt valve at about 500 kc/s. Frequency was stabilised to 0·01 per cent. for a capacity-variation corresponding to 0·1 per cent. frequency-change, and frequency-change due to anode-voltage reduced from over 0·02 per cent. to 0·005 per cent. These were obtained with a power-output of over 100 watts. At frequencies higher than one

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megacycles per second, screen-grid valves are recommended, to lessen the "feed-back" of power to the grid-circuit through the grid-anode capacity. In this way a power-output of 100 watts was obtained at 3 megacycles per second without excessive heating of the crystal.

It is pointed out that the stabilised range of frequency (with respect to capacity) is less than that of ordinary quartz-controlled circuits, but is considered to be adequate for many practical cases.

The third part of the paper deals with two circuits in which it is sought to obtain automatic monitoring of the frequency of an oscillator of relatively large power by reference to the frequency of a small-power master-oscillator of some stabilised form, or to the resonant frequency of a quiescent circuit of low decrement (*e.g.*, a quartz crystal). In one circuit described, a mechanical arrangement is used in which the variation of the heterodyne-frequency due to interaction with the master-oscillator causes a change of capacity of appropriate sense in the power-oscillator. In such a circuit at 100 kc/s. the frequency-variations due to antenna swaying were reduced to about one-tenth, and the initial frequency-drift similarly reduced from 0·1 per cent. to 0·01 per cent. It is claimed that applied to a power-transmitter with only one stage of amplification, a stability of 0·01 per cent. could be obtained in this way with a quartz-controlled small-power oscillator temperature-controlled to 1° C.

In another form of circuit described, an auxiliary power-valve is associated with the inductance of the oscillator in such a way that it acts as a pure reactance, of amount depending on the grid-bias of the auxiliary valve. A quartz-resonator is coupled to the main oscillator in such a way as to give a response which is critically dependent on frequency, both as to magnitude and sense (*i.e.*, variation about a fixed point on the side of the crystal resonance-curve is used). This variation produces a corresponding voltage-variation in the anode-circuit of a valve operated by the crystal, and this voltage-variation is applied to the grid of the auxiliary valve referred to above in such a way as to oppose the frequency-change.

As applied to a small-power valve (frequency not stated) it is shown that frequency is stabilised to 0·01 per cent. for a capacity-change corresponding to a frequency-variation of 0·1 per cent., and that similar stabilisation is obtained with respect to supply-voltages.

Comments

1. The special circuit-arrangements described, particularly the first, appear to be very useful for minimising frequency-variations due to valve changes alone. The principal application would seem to be to master-oscillators. Though the stabilising resistance will vary somewhat with frequency, it seems to be applicable to cases where frequency-changes or frequency-adjustments are

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required. It appears, in fact, to be a simple means of obtaining similar constancy to that given by the Llewellyn circuits.

2. The new form of crystal-controlled circuit raises a number of points of theoretical interest with important practical application. The circuit appears to be essentially similar to that described by Mabboux (abstract No. 15) for frequency control by a tuning-fork.

3. The automatic-monitoring schemes described are considered to be of special importance, not only in themselves, but as representative of the whole idea of automatic monitoring, an idea which is as yet comparatively unexplored. The general aspects of this matter are considered more fully in section 7. As far as concerns the particular circuits described, the most important comment that arises is that in neither of the circuits is the sense of the controlling-force made independent of the amplitude of the transmitter oscillation. This independence may be an important desideratum in practice, as, in the absence of it, perturbation of frequency may arise as a consequence of keying.

The paper as a whole can be described as a most useful and suggestive contribution to the subject.

18.1. **Automatic Frequency-Regulator for Short-Wave Transmitting Stations.** ANON.: *Bull. de la S.F.R.*, January, 1928

Abstract

This describes an automatic-monitoring system which is actually in operation in the short-wave stations of the S.F.R. Briefly, the principle is as follows. The heterodyne-response due to interaction between the transmission and a small-power constant-frequency oscillator is amplified by a system having a fairly steep frequency-characteristic, and is then rectified, further amplified as d.c., and passed through a winding on an iron core, which is also linked with part of the main inductance of the transmitter. The response also operates an "electromagnetic condenser" (apparently an electromagnet-operated arrangement of which the capacity varies with the operating current). The intention is that a change of heterodyne-frequency one way or the other will cause an appropriate change in the inductance through the variation of permeability of the iron core by reason of the change of steady flux-density. The electromagnetic condenser (of high time-constant) is intended to deal similarly with slow drift or other slow changes of frequency.

The paper gives a detailed description of the component elements of the system. It is claimed that the frequency-regulation obtained is better than 1 in 100,000.

Comments

The system shows great ingenuity both in principle and in detail, and is very interesting as an actual example of an automatic-

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ally-monitored scheme for short-wave transmissions (see abstract No. 17).

The following appear to be features which might lead to difficulties in operation :—

(1) Any small change in amplitude in either the transmission or the independent master-oscillator would apparently act in the same way as a change in frequency in either.

(2 The iron core, consisting of nickel-iron laminations, with mica separation, has to be oil-cooled. Temperature-variations in this core will produce permeability-variation and consequent frequency-changes.

It may also be pointed out that the whole assembly of controlling mechanism is somewhat elaborate.

18.2. **The Single Side-band System Applied to Short-Wave Telephone Links.** A. H. REEVES: *J.I.E.E.*, 1933. *See also* **The Single Side-band System Applied to Short Wave-Lengths.** A. H. REEVES: *Electrical Communication*, 1931, **10**, 3-19

Abstract

The feature of the above publications which is relevant to the present subject is a system of automatic monitoring of a local oscillator by reference to a received "pilot" signal. The local oscillator is set to give a heterodyne beat-frequency of 4 kilocycles per second with the incoming frequency. This beat-frequency is amplified and made to excite two circuits which are tuned respectively to 3·8 and 4·2 kilocycles per second. These circuits are associated with rectifiers in such a way that the rectified responses are in opposition, giving zero resultant output potential-difference when the input frequency is exactly 4 kilocycles per second (midway between the two tuning points). A frequency-aberration of the local oscillator relative to the incoming signal will therefore give a rectified potential-difference response the sense of which depends on that of the aberration. This potential-difference is applied to the grid of an auxiliary valve in the anode-circuit of which is an inductance coupled to that of the local oscillator, and so arranged that the variation of the effect of this inductance on the oscillator-frequency with any change of the auxiliary valve grid-bias is such as to tend to nullify the change. It is claimed that if the frequency is allowed to vary by 5–6 kc/s., the local oscillator will follow the variations to within 20 cycles per second.

Comments

The above circuit could clearly be used to control a transmitter by reference to a local oscillator of stable frequency. As applied to a keyed transmission it would, however, have the disdvantage

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that if the transmitter drifts appreciably from its nominal frequency, the correction by the monitoring circuit has to cover the whole frequency-aberration each time the key is pressed. This drawback would not, however, apply to a two-stage transmitter in which the drive was continuous.

19. Frequency-Stability of a Self-maintained Transmitter. V. I. AGUIEFF and L. A. KOPYTIN: *Doc. du C.C.I.R. for Second Meeting*, 1931, 147-156

Abstract

The above appears (without title) as an appendix to the contribution on frequency-stabilisation from the U.R.S.S. It describes, essentially, the application of the principles discussed in abstract No. 3 (also taken from the documents of the C.C.I.R.) to a 1 kw. transmitter of conventional type (Meissner circuit with tuned anode coil) at about 500 kc/s. It is shown by variation of the component elements, particularly L/C ratio and coupling, that the appropriate design of the circuit does in fact lead to a high degree of frequency-stability in respect of variation of supply-voltages. In the final disposition the circuit showed a stability better than 0·01 per cent. with respect to deliberately imposed supply-voltage changes of degree greater than those likely to occur in practice.

It was found, however, that an initial drift of about 0·1 per cent. occurred in the first two hours (a steady fall of frequency). This was traced to temperature-changes in the circuit. When a system of cooling was applied to the capacitance and inductance, this drift was eliminated. It was also eliminated more simply by using an inductance of 2½ cm. diameter copper tube and an air condenser with large cooling surface and low resistance.

Comments

This is essentially a practical contribution to the subject. The inference appears to be that an entirely conventional self-excited transmitter-circuit can be made to give a stability of the order of 0·01 per cent. by care in the detail of the design. It is interesting to note that in the case examined there is definite evidence that the initial drift was not due to valve-changes but to circuit-changes. On the other hand, operation at higher frequencies, with correspondingly smaller tuning capacity, would probably have shown a valve-effect (*i.e.*, change of interelectrode-capacitance) as one component of the drift.

It must be pointed out that changes due to load-variation (*e.g.*, aerial-capacity, etc.) are not included in the above investigation, but the work is in any case applicable to the stabilisation of a high-power master-oscillator.

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20. **Inductances of High Permanence.** W. H. F. GRIFFITHS: *J.S.I.*, 1929, **6**, 354-357

Abstract

It is stated that single-layered coils wound on ebonite are likely to be unstable with respect to thermal history because ebonite is not in general cyclic in this respect. Moreover, if the winding is under tension the wire may be permanently stretched by the greater expansion of the ebonite. Bakelite is better than ebonite in this matter as its coefficient of expansion is similar to that of copper.

The construction described consists of an ebonite cylinder mounted between bakelite end plates, the cylinder being segmented so that its diameter is determined by the fixing to the end plates. A loaded ebonite is used for the cylindrical part, and is of such length and coefficient that the decrease of inductance due to its expansion balances the increase due to the diametral expansion of the copper wire and bakelite end plates.

A similar multilayer construction is also described.

Comments

The coils described have apparently proved satisfactory in respect of thermal changes in the application for which they were designed, *i.e.*, as wavemeter coils. They would presumably be equally satisfactory for small-power master-oscillators, but do not appear applicable to large-power transmitter-circuits.

21. **A Valve Generator of Very Constant Frequency.** P. DAVID: *Comptes Rendus*, 1931, **192**, 1209-1210, and *Rev. Gen. d'Elect.*, 1931, **29**, 230D. *Fr. Pat.* 700498

Abstract

A four-electrode (space-charge grid) valve is used, the tuned anode-circuit being inductively coupled to *both* grids by way of variable tappings on the grid coupling coil. Stability of frequency of the order of 5 parts in a million for 15 per cent. changes of supply-voltage are claimed.

In the C.R. account it is said that no complete theory has yet been obtained, but in the patent specification it is said that the stability is due to complete suppression of grid-current and to the restriction of anode-current to impulsive flow of short duration suitably phased.

Comments

If the mechanism of the maintenance is by impulsive supply of energy, as suggested, the circuit is of interest as providing the electrical analogy of maintenance of a pendulum by impulses

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delivered at the moment of zero displacement (*c.f.* Edgworth, abstract No. 2.1).

Very little detailed information is given, but the impulsive type of maintenance indicated, though probably inapplicable to transmitters, should have good features as applied to master-oscillators.

22. The Variation of the Resistance and Interelectrode Capacitances of Thermonic Valves with Frequency.

L. HARTSHORN: *Wireless Engineer*, 1931, **8**, 413-421

Abstract

It is shown that the interelectrode capacitances of a thermonic valve with the filament emitting may be expected to differ from their values when the filament is cold, on account of the presence of the space-charge in the former case. At very high frequencies the values will tend to coincide, but at low frequencies the space-charge affected value may be 7 per cent. or so higher than the other. As a function of frequency the variation from the higher to the lower value is likely to occur mainly in the range 10^6 to 10^8 cycles per second.

The effective resistance should show a similar variation as between low and high frequency values, the variation being, by theory, of appreciable extent (12 per cent. or so) and occurring mainly in the same frequency range.

Experimental results at audio frequencies give qualitative confirmation, but the capacitance increase at these frequencies, as compared with the cold or geometrical value, is greater than that predicted.

Comments

The interest of this paper from the point of view of constant-frequency oscillators is the demonstration that interelectrode capacitance may be expected to vary with the space-charge distribution, particularly at frequencies in the region of 10^6 cycles per second upwards. The method of analysis adopted does not permit of any quantitative estimation of the effect as applied to such variations of space-charge distribution as may occur in an oscillator, but it would appear that variations of the order of a micro-microfarad are not impossible. Such variations would give rise to very appreciable frequency-changes in certain forms of oscillatory circuit.

23. An Oscillator having a Linear Operating Characteristic.

L. B. ARGUIMBAU: *P.I.R.E.*, **21** (1933), 14-28

Abstract

It is pointed out that in any ordinary back-coupled valve maintained oscillator, " no non-critical stable equilibrium can be established unless the tube is permitted to operate over a thoroughly

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non-linear region. . . . If it were possible to set up an oscillator whose damping increased rapidly with increases in amplitude, preferably linearly at low intensities . . . the tube would function linearly without distortional troubles."

A circuit intended to realise this condition is described. It is normal as to the tuned anode-circuit and grid-winding, but an additional winding, coupled to these, excites a linear rectifier of which the d.c. response is applied in appropriate sense as an additional grid-bias of the oscillator valve. The result is that the anode-voltage variation is restricted to a certain small and controllable equilibrium-value, since any increase of this amplitude will cause a further reduction of control-grid voltage with a consequent tendency to return to the initial equilibrium-amplitude. It is claimed that a truly linear variation of amplitude with an applied grid-circuit modulating-voltage will result.

It is pointed out that the circuit differs only from the normal grid-leak and condenser type, in which the grid-bias is also a function of oscillation-amplitude, in the nature of the functional relationship between grid-bias and amplitude, which is not nearly as closely linear in the grid-leak oscillator as in the new circuit.

An analysis, including non-linear and " dynamic " effects (*i.e.*, an effect in modulation comparable with " side-band cutting " in reception) is given. It is shown that a high ratio of grid-bias change to oscillation-amplitude change is the most important factor.

Comments

It is stated that " computations show that such an oscillator has high frequency-stability," but no measured results are given. The idea would seem to have useful possibilities, which are further discussed in section 2.4 of Part I.

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DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH

RADIO RESEARCH

SPECIAL REPORT No. 14

MAGNETIC MATERIALS AT RADIO FREQUENCIES

A CRITICAL SURVEY OF PRESENT KNOWLEDGE

BY

F. M. COLEBROOK, B.Sc., D.I.C., A.C.G.I.

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PREFATORY NOTE

IT is well known that iron and iron alloys in a suitable state of sub-division into thin laminations or in the form of powder-compositions, have properties of great value in connection with line-telegraphy and telephony, and the behaviour and characteristics of such materials at frequencies up to the highest involved in line-telephony have been very widely studied.

It has recently been found possible to make ferro-magnetic-powder compositions in which the size, distribution and insulation of the particles are such that the advantages of a magnetic core-material are retained to a useful degree up to radio frequencies of the order of a million cycles per second. At present, however, there is comparatively little exact knowledge of the behaviour and characteristics of such materials at radio frequencies, and the subject has therefore been considered by a sub-committee of the Radio Research Board of the Department of Scientific and Industrial Research.

On behalf of this sub-committee a survey has been made of existing knowledge of magnetic materials and compositions at radio frequencies, including also such information relating to audio-frequency behaviour as is likely to prove useful as a guide in any future investigation of the subject at radio frequencies.

In view of the general interest and the possible importance of this recent development in radio-frequency technique, it has been decided by the Radio Research Board that a useful purpose might be served by the present publication of the survey.

A. G. LEE
(*Chairman Radio Research Board*).

January, 1934.

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MAGNETIC MATERIALS AT RADIO FREQUENCIES

A CRITICAL SURVEY OF PRESENT KNOWLEDGE

1. OBJECT AND SCOPE OF SURVEY

THE object of the survey is to summarise existing knowledge, theoretical and experimental, of the behaviour of magnetic materials at radio frequencies. The practical value of such materials lies in the possibility of obtaining lower ratios of resistance to inductance than can otherwise be obtained, in a given space, in coils used at radio frequencies, together with improved facility in screening, and more effective couplings between different coil-windings. The survey is mainly directed to this practical objective. No range of radio frequency has been excluded, but particular attention has been directed to frequencies higher than about 100 kilocycles per second and to iron-powder core-materials suitable for use at these higher frequencies.

A bibliography is appended. This is not intended to be fully comprehensive, but is a selection conditioned by the practical objective indicated above.

2. CLASSIFICATION OF THE CHARACTERISTICS OF COMPOSITE MAGNETIC MATERIAL

Magnetic material as used for cores of coils carrying radio-frequency currents is invariably a constructed mass consisting partly of magnetic metal and partly of some insulating material. The behaviour of any given mass of such material in a radio-frequency field depends upon elements which are, respectively (1) characteristic of the constituents of the material, and (2), for given constituents, determined by their individual and relative configurations and the configuration of the mass as a whole.

2.1. MATERIAL-CHARACTERISTICS

2.1.1. *Magnetic constituent*

The principal material-characteristic of the magnetic constituent is permeability. Hysteresis loss is also of this kind, and there is

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another loss factor which is a material-characteristic but which only appears when the material is subject to a varying field. The German name for this element is " Nachwirkung." There seems to be no generally accepted English name, but " retardation-loss" conveys the sense of the German term. The only remaining material-characteristic involved is volume-resistivity.

2.1.2. *Insulating constituent*

The most important material-characteristic of the insulating constituent is resistivity. There is, however, a second characteristic which may also be of appreciable significance, namely, the power-factor of the material considered as a dielectric.

2.2. Configuration-Characteristics

A mass consisting partly of magnetic metal and partly of an insulating material will have an effective permeability, different from that of the magnetic metal, and not, in general isotropic. This effective permeability is therefore a configuration-characteristic. The same applies to effective resistivity, which also will not in general be isotropic.

The electro-motive-forces induced in magnetic material when it is subjected to a varying electric field give rise to eddy-currents. These produce a loss of power depending on the individual and relative configurations of the constituents. Eddy-current loss is thus a configuration-characteristic. The counter magneto-motive-forces due to the eddy currents have the effect of a reduction in permeability to a new apparent value. Apparent permeability in this sense, depending both on effective permeability and on the eddy-current distribution, is thus a configuration-characteristic. The total hysteresis-loss (and, to a less extent, retardation loss) will also be affected by the resultant field distribution and are thus configuration-characteristics dependent on material-characteristics. Similar considerations apply to dielectric loss.

The above classification of the actual, effective, and apparent characteristics of magnetic materials may seem somewhat elaborate, but a search of the literature has shown that in some cases experimental results have been wrongly interpreted, or not presented in the most informative manner, on account of the complicated interaction of the various characteristics, and the above attempt at clarifying the fundamental ideas has therefore been thought desirable.

The various material and configuration-characteristics enumerated above will now be considered separately in relation to radio-frequency operation.

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3. PERMEABILITY

3.1. Variation of Actual Permeability with Frequency

Some early measurements on the variation of permeability of magnetic metal with frequency showed critical variation in the neighbourhood of three megacycles per second. In consequence much attention has been given to this matter by a number of workers,[10–14] and the results of these investigations can be summarised by the statement that the initial permeability μ_0 in the Rayleigh region of linear permeability variation, *i.e.*,

$$\mu = \mu_0 + \alpha H$$

is substantially independent of frequency at least up to 300 megacycles per second. Max Wien finds, however,[12] that α decreases somewhat as frequency increases (range investigated 0·1 to 10 megacycles per second), and that the Rayleigh equation holds to higher values of H at high than at low frequencies. Thus, for example, the upper limit of H for linearity may be, for a given material, about 3 gauss at low frequencies or for steady fields, increasing up to 8–9 gauss at high radio frequencies. The decrease of α with increase of frequency will of course imply a small decrease in μ with frequency for larger values of H. The magnetising force will usually be fairly small, however, in radio-frequency application. (It is not explicitly stated in Max Wien's paper, which is of the nature of a preliminary announcement, that his data are of actual, as distinct from apparent, permeability, but he has kindly confirmed that this is so in a letter to the present writer. He points out, however, that his allowance for the effect of eddy-currents depends upon an approximated theory, the validity of which is being further investigated).

3.2. Effective Permeability of Iron-Powder Agglomerates

The introduction of iron-powder cores lead to a number of theoretical and experimental investigations[6, 7, 8] of the dependence of μ_{eff} (effective permeability) on μ (actual particle permeability) and on the relative configurations of the particles and the binding or insulating medium. Howe,[22] for example, assuming a uniform distribution of uniform cubic particles arrives at the formula

$$\mu_{eff} = \frac{\mu - \frac{2}{3}\delta(\mu - 1)}{1 + \frac{\delta}{3}(\mu - 1)},$$

where δ is the fractional volume of the insulation. This formula gives an increase of μ_{eff} with μ, but the rate of increase becomes

* These numbers refer to the bibliography on p. 21.

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very slow for $\mu > 100$–200. Thus, for $\delta = 0{\cdot}1$, $\mu = 200$, $\mu_{eff} = 23$, and the limit of μ_{eff} is only 28 when $\mu \to \infty$. All the effective permeability formulæ, some of which are more elaborate than Howe's in consequence of the inclusion of anisotropic particle distributions, agree in the main conclusion that the configuration and distribution of the particles are very much more important than particle-permeability in determining the effective permeability. They agree also that a figure of about 80 is the highest practically attainable, even with particles of the highest permeability. This is apparently about the highest value that is usually reached in actual powder-compositions,[29] though there is a commercial product of this kind for which a permeability of over 100 is claimed.

In core-materials intended for use at radio frequencies, however, the ratio of effective permeability to the core-losses is the significant quantity. This, as considered in more detail below, imposes restrictive limits on particle size and spacing and appears to preclude the attainment of high effective permeabilities in such materials. A. SCHNEIDER, for example,[23] states that 18 is the highest "economic permeability" for the powder-core material developed by VOGT for use at radio frequencies, and in the existing types of iron-powder radio-frequency core-materials the effective permeability is usually even lower than this, in some cases as low as 4 or 5.

The loss in permeability is presumably due for the most part to the gaps lying in planes normal to the flux. HOWE points out that insulated iron wire of sufficient fineness, lying parallel to the flux, would be more suitable in this respect, but is probably not practicable. SCHNEIDER states that this possibility had been considered by VOGT but found to be impracticable.

Constancy of effective permeability with respect to induction is of importance in relation to the radio-frequency application of iron-powder cores. DEUTSCHMANN[5] discusses a "non-linearity" factor, based on the assumption that the so-called Rayleigh-region is similar to that for d.c. operation. As stated in 3·1, however, the μ-H relationship is not independent of frequency at radio frequencies, the variation being in the direction of greater constancy of μ.[13] SHACKLETON and BARBER[29] show that the d.c. permeability of powdered permalloy is constant up to about B = 100, and does not vary by more than 10 per cent. up to B = 5,000. Cores of iron powder electrolytically produced are shown by them to have a sensibly constant permeability up to about 60 for B. No data have been found for the d.c. permeability of any other powder-compositions.

Constancy of effective permeability with respect to time (secular permanence), and constancy with respect to magnetic shock or overloading (magnetic stability) are also factors of obvious practical importance. SHACKLETON and BARBER,[29] and MARRIS, POLGREEN

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and WILLIAMS[31] show that the behaviour of the more recent types of powder-core telephone-cable loading-coil is quite satisfactory in these respects, the permanent inductance-variation due to magnetic shock not exceeding 0·5 per cent. in an extreme case. There seems to be no published data on this aspect of core-materials for use at radio frequencies but, assuming adequate stability, the relatively low effective permeabilities of these materials may be expected to give a correspondingly low susceptibility to magnetic shock.

Constancy of effective permeability with respect to temperature is a desirable feature, particularly in radio-frequency applications. In itself, the subdivision of the ferro-magnetic constituent of a powder-core would be expected to result in a temperature coefficient of permeability lower than that of the ferro-magnetic material itself, but there is at present practically no published information on the point. The writer has reason to believe that the temperature coefficients of inductance of coils wound upon cores of iron-powder compositions may be positive or negative and generally of the order of a few parts in ten thousand. Where there is appreciable variation of permeability with temperature, it is particularly important in radio-frequency applications that such variation shall be regular and cyclic. On this point also there appears to be no published information.

3.3. APPARENT PERMEABILITY IN RADIO-FREQUENCY OPERATION

The counter magneto-motive-forces due to eddy-currents result in a reduction of the total flux through magnetic material in a radio-frequency field, as compared with that for an equal field of zero or low frequency. In contrast to " skin-effect " in a.c. distribution, induction is everywhere reduced by the action of eddy-currents, and not merely re-distributed. There is in consequence a reduction of permeability to a new value, which, for distinction from effective value (Section 3.2) will be referred to as an apparent value. The subject is included in LATOUR'S[3] analysis of eddy-current formation, and a very clear picture of the effect in iron laminations is given by WOLMAN.[4] For a sheet of thickness t cms. resistivity ρ (in E.M. units), permeability μ, there is a certain " bounding frequency " (" grenz frequenz ") given by

$$f_g = \frac{\rho}{\pi^2 t^2 \mu}$$

(*cf.* uniform-flux formula in 4·3, which is of the same form). Taking as an example $\mu = 250$, $\rho = 5 \times 10^4$, $t = 2{\cdot}5 \times 10^{-3}$ (1 mil), $f_g = 3 \times 10^6$.

The practical significance of the bounding frequency, so defined, is that for all frequencies up to the bounding frequency the apparent permeability is substantially constant and equal to the actual permeability, while for frequencies higher than the bounding

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frequency the apparent permeability decreases approximately as the inverse square root of the frequency.

WOLMAN gives experimental confirmation of his analysis, but his frequency-range is not specified. ALEXANDERSON states that the variation of apparent permeability in the measurements described by him[15] is consistent with the theory of the skin-effect. Calculations based on his data confirm the theoretical variation as $f^{-\frac{1}{2}}$.

It should be noted that in the case of iron-powder cores, there may be mass-conductivity effects on permeability, superimposed on those of particle-conductivity. In such cases there might be substantially uniform flux-distribution in the individual particles, but some degree of non-uniformity of flux-distribution through the mass. The effect would be conditioned by the relevant mass-resistivity, which is in general very high compared with that of the particles.

4. HYSTERESIS

4.1. DEPENDENCE ON MECHANICAL AND THERMAL HISTORY

It is known that the hysteretic vehaviour of magnetic metal depends on its thermal and mechanical history. There appear to be no published data showing specifically the effect of fine subdivision on hysteretic behaviour. It is found, however, that the hysteretic loss in powdered permalloy can be minimised by suitable heat-treatment, as in the solid material. There is no evidence of any very abnormal hysteretic behaviour due to fine subdivision.

4.2. EFFECTIVE HYSTERESIS IN IRON-POWDER AGGLOMERATES

Apart from a possible effect of fine subdivision in itself, the d.c. or low-frequency hysteretic behaviour of compositions of powdered magnetic metals will be a configuration-characteristic, there being, so to speak, a dilution of hysteretic loss consequent on the dilution of the magnetic material.[5] The particle flux-density will, of course, be higher than the effective flux-density of the mass, but very much lower than that for a similar mass of the pure magnetic material. As DEUTSCHMANN points out,[5] the d.c. or low-frequency behaviour of a closed ring of such material will be comparable in respect of flux-density and hysteresis with that of a ring of the magnetic material alone, there being an air-gap in the ring. SPEED and ELMEN make the same observation.[28]

4.3. VARIATION OF HYSTERESIS AND HYSTERESIS LOSS WITH FREQUENCY

It appears to be fairly well established that at frequencies below about 1,000 cycles per second there is a constant hysteretic loss

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per cycle for a given flux-density.[17] On this basis W. DEUTSCHMANN has given an analysis of the effect of hysteresis on the behaviour of telephone-cable loading-coils, it being assumed that operation is confined to the so-called Rayleigh-region, for which

$$B = \mu_0 H + \nu H^2$$

For maximum or amplitude values B_m and H_m, the hysteretic loop is given by

$$B = (\mu_0 + 2\nu H_m) H \pm \nu (H^2 - H_m^2)$$

and the hysteresis-loss in ergs per cycle per cubic cm. (*i.e.*, the area of the hysteretic loop divided by 4π) by

$$W_h = \frac{2\nu}{3\pi} H_m^3$$

The corresponding hysteresis-loss component of the resistance of a coil of inductance L wound upon a core of such material is, at frequency f cycles per second, approximately

$$R_h = \frac{16\nu}{3\mu_0^2} f B_m L.$$

In terms of the familiar Steinmetz formula

$$W_h = \eta B_m^x$$

with $x = 3$ this becomes

$$R_h = 8\pi\eta \mu_0 f B_m L$$

The behaviour of iron-powder core materials appears to be consistent with this formulation at audio frequencies[31] but the observations recorded in section 3·1 and the further points mentioned below make it doubtful whether it will adequately represent the hysteresis-behaviour of such materials at radio frequencies.

The actual evaluation of the hysteretic component of loss at radio frequencies or even at high audio frequencies is complicated by eddy-current formation and skin-effect, with a consequent non-uniform flux-distribution, leading to a frequency-variation of apparent hysteretic loss not easily distinguishable from that of the eddy-current loss. The reliability of such loss-analysis is therefore limited to that of the calculation of the eddy-current lossses. With this reservation, the evidence tends to show that at frequencies higher than a few thousand cycles per second, the hysteresis-loss per cycle increases with frequency.

Thus NEUMANN finds[19] a 60 per cent. increase at 1,500 cycles per second over the static value for a 50 per cent. iron-nickel alloy. BRASEY, in a very thorough and convincing set of measurements on iron wire of 0·24 mm. diameter,[17] finds an increase of about 15 per cent. at 10,000 cycles per second. It appears, therefore,

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that apart from the eddy-current complication of the hysteretic effects, the loss per cycle for a given flux-density may vary considerably with frequency, and may be very much higher at, say, 10^6 cycles per second than the static value. Thus, the relative importance of hysteretic and eddy-current loss at radio frequencies cannot at present be taken as known, though it is very probable that the latter predominates.

At radio and high audio frequencies the counter magneto-motive-forces due to eddy-currents produce a decrease of flux in the interior of the magnetic material (laminæ or particles). For a given applied M.M.F. this will in itself result in a decrease of hysteretic loss, but for a given mean flux-density, in an increase of hysteretic loss.[1] (A given total flux will produce minimum hysteretic loss in a uniform distribution.)

A detailed analysis of the flux-distribution and its effect on hysteretic loss is given by LATOUR[3] but it is based on simplifying assumptions (loss per cycle for given density independent of frequency: $W_h = \eta B^2$) which make it of doubtful validity at radio frequencies.

It appears probable (see sect. 7) that most practical applications of iron at radio frequencies require as an economic condition a fineness of subdivision or lamination such that the flux-distribution remains substantially uniform. Under these conditions any variation of hysteresis-loss will be mainly that due to a variation with frequency of the loss per cycle, uncomplicated by any secondary effects of eddy-currents on flux-distribution.

5. RETARDATION LOSS

As hysteresis-loss is proportional to some power of the induction, lines representing total loss in a.c. operation, less losses due to eddy-currents, plotted against induction, should pass through the origin. It has been found,[5] however, comparatively recently, that this is not the case in actual fact. The additional loss term, apparently independent of B, thus revealed, has been called, in German, " Nachwirkung "—the sense being not " after-effect " as that would be understood in English, but " retardation." The name is apparently due to the theory that the loss arises from a lag of B behind H due to a kind of magnetic viscosity.

DEUTSCHMANN describes the coil resistance component R_n due to this cause as being proportional only to L (*i.e.*, μ) and to f.[5] Comparatively little has been published on this subject, and nothing at all relating to radio-frequency operation. DEUTSCHMANN shows it as being comparable with eddy-current loss in laminated and powder cores at audio frequencies, but it will presumably increase less rapidly than eddy-current loss at higher frequencies. At present, however, its significance at radio-frequencies cannot be assessed.

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There is no obvious reason to suppose that it can be minimised by fine subdivision of the magnetic material.

6. RESISTIVITY

In the case of composite cores—laminations or powder—two distinct resistivities are involved, namely, that of the magnetic constituent and the mass-resistivity. This latter may not be isotropic, but the resistivity along paths normal to the flux will be the important factor.

The mass-resistivity will in general be high compared with the particle-resistivity. Thus, in the powdered permalloy cores it is given as 1–20 ohm-cms., as compared with something of the order of microhms for the metal itself. It is, however, of greater practical significance than is suggested by this disparity, since very much larger induced e.m.f.'s are effective in the mass than in the particles.

There is very little published information on the mass-resistivity of existing iron-powder core-materials for use at radio frequencies. In one case[27] the resistivity is given as 50 ohm-cms. In another type[23] a very high mass-resistivity is attained, probably at the cost of some sacrifice in effective permeability, by a lamination of the mass by means of fine tissue paper.

A practical point of some interest in emphasised by SPOONER.[1] On a basis of equal M.M.F., eddy-current losses for uniform flux distribution and for a surface flux distribution are respectively as f^2, B^2, ρ^{-1} and $f^{\frac{1}{2}}$ $B^{\frac{1}{2}}$ $\rho^{\frac{1}{2}}$. In the second case, the losses will increase with resistivity.

7. EDDY-CURRENT LOSSES AT RADIO FREQUENCIES

The theory of the formation of eddy-currents in magnetic material in alternating fields, and the consequent decrease of induction within the material, has been very fully explored. LATOUR'S analysis[3] is very comprehensive and seeks to include the effects of hysteresis.

HOWE has given an elegant and simple analysis for iron laminations, based on transmission-line theory,[32] leading to the conclusion that the power-loss due to eddy-currents in laminæ of thickness t will be proportional to $B^2t^2f^2\rho^{-1}$ at low frequencies, and will tend to proportionality to $B^2t\,f^{1\cdot5}\,\mu^{-\frac{1}{2}}\,\rho^{-\frac{1}{2}}$ at higher frequencies. These results are in agreement with those of LATOUR.

ROSENBERG'S theory[2] is very simple in form but is practically restricted to the limiting cases where the flux is confined to the surfaces of the magnetic elements of the material. WOLMAN'S analysis[4] is presented in a very clear and graphic form. Though specifically related to iron laminæ, his main results appear to have some general application.

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The value of the analyses of this subject appears to be qualitative only. Comparison of measured and calculated values leads SPOONER[1] to the conclusion that the uncertainties in the various formulæ are such that " reliable high-frequency loss values can be obtained only by test."

The matter is complicated by the interaction of all the various characteristics of the material at high frequencies, but some little simplification can be achieved by a classification based on WOLMAN'S conception of a " bounding frequency " (grenz frequenz), characteristic of any given configuration. For a lamina of thickness t, for example, the bounding frequency f_g is given by

$$f_g = \frac{\rho}{\pi^2 t^2 \mu}.$$

The physical significance of the bounding frequency has already been indicated (section 3.3). Its utility as a basis for classification is that the reduction of flux and therefore of apparent μ by eddy-current formation is, as a function of frequency, fairly slow up to the bounding frequency and fairly rapid thereafter.

The nature of the variation with frequency is most conveniently depicted (as by WOLMAN)[4] in vector form, *i.e.*, as a variation of vector or complex permeability. Neglecting for the moment copper and hysteresis losses, the potential difference v across an iron-cored inductance L at frequency $\omega/2\pi$ is given by

$$v = \mathrm{L}\, di/dt.$$

Putting $\mathrm{L} = k\mu$, and considering v and i as sinoidal gives, in vector form

$$v = k\mu\, j\omega i.$$

This will apply at very low frequencies. At higher frequencies the power-loss due to eddy-current formation can be represented by associating a negative imaginary component with μ, *i.e.*,

$$\begin{aligned} v &= k\, j\omega\, (\mu - jr)\, i \\ &= (k\omega\, r + [k\omega\, j\mu)\, i \\ &= (\mathrm{R} + j\omega\mathrm{L})\, i \end{aligned}$$

The power-factor of the coil under these conditions is $\cos\phi$ where

$$\tan\phi = \frac{\mathrm{R}}{\omega\mathrm{L}} = \frac{r}{\mu}$$

We have thus the conception of a power-factor for the permeable medium, similar to that for a dielectric, with a vector or complex apparent permeability.

$$\mu_c = |\mu_c|\, \varepsilon^{j\phi} = \mu - jr.$$

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WOLMAN shows that up to $f = f_g$, μ_c has an approximately circular locus as a function of f, *i.e.*, putting μ_0 for the initial permeability,

$$\mu_c = \frac{\mu_0}{2}(1 + \epsilon^{2j\theta}),$$

θ being a function of frequency. There follows an intermediate condition up to $f \simeq 2 f_g$ and thereafter the variation is represented by

$$\mu_c = \frac{\text{const.}}{\sqrt{f}} \mu_0 \epsilon^{-j\pi/4}$$

Thus, as far as eddy-current loss is concerned, there is a limiting phase angle of 45° as the frequency increases indefinitely, giving a power factor of 70 per cent. (This generalisation was first published by J. J. THOMSON). The effect of hysteresis and copper-loss will be a small further reduction of θ below $\pi/4$. Thus, ALEXANDERSON records power-factors of 70–80 per cent. for f = 100–200 kilocycles per second for 3 mil soft iron.[15]

WOLMAN gives experimental confirmation of his analysis, though not in detail. The recorded measurements of losses and of apparent permeability at radio frequencies appear to be in general agreement with the above description.[15, 16, 18, 20, 21]

ALEXANDERSON states explicitly that in his measurements the variation of apparent permeability was that which would be expected from the theory of the skin-effect, actual permeability remaining substantially constant. A calculation based on his data gives a fairly constant value of $B\sqrt{f}$ for a given H for values of f from 60–200 kilocycles per second.

As far as radio-frequency application is concerned, only the first region of behaviour, where f is small compared with f_g (the bounding frequency) is of any practical interest. Assuming the validity of the above analysis at radio frequencies we thus arrive at a practical criterion of utility for magnetic material at these frequencies. Its configuration must be such as to give a bounding frequency high compared with the frequency of operation, under which conditions the power-factor can be kept reasonably low—of the order of 1 per cent. or so, with eddy-current losses proportional to f^2 and B^2 and with an apparent permeability which does not very rapidly with frequency nor differ greatly from its low-frequency effective value.

With the experimental data at present available it cannot be stated how far the above discussion is actually applicable to existing forms of magnetic material at radio frequencies, but one significant feature may be noted. The circular variation of vector permeability referred to above can be represented[4] as the effect of a constant resistance in parallel with the inductance. The tuned rejector-circuit impedance of such a coil, tuned by a parallel capacitance,

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would be constant with respect to frequency, and equal to the shunt resistance. The measurements on certain coils having toroidal cores of an iron-powder composition (see section 10), give results of this character. To this extent the observed behaviour is consistent with WOLMAN'S analysis, which may therefore be useful as a guide in further investigation.

The conceptions of complex permeability and of the power-factor of a permeable medium are considered more fully in relation to the total losses in the core, in section 9.1.

8. DIELECTRIC LOSSES IN IRON-POWDER CORE-MATERIALS

An article by SCHNEIDER[23] contains the passage :—" There are produced in the core not only galvanic eddy-currents, flowing in each individual particle and extending over a group of particles being in contact with one another, but also eddy-currents due to capacitive coupling . . . Each two magnetic particles, which are separated from each other by the insulating skin, form a little condenser. . . . Thus a capacitive coupling is formed and the eddy-current, which cannot flow in the usual galvanic way due to the insulating skin, passes as a dielectric current from particle to particle." It is stated in the article that considerable dielectric loss is caused in this way, and that the loss is minimised by the laminated and striated configuration of the core-material described in the article. It is implied that the conclusions with regard to this loss-component are based on experiment, but the experimental evidence has not yet been published. The conclusions have been criticised by HOWE[34] and NISSEN[35] who show by analysis of a simplified particle-dielectric configuration that the loss-component associated with such dielectric eddy-currents is not likely to be appreciable, even though, as pointed out by HOWE in a later note,[36] the presence of the conducting particles in the insulating matrix will result in a higher effective dielectric constant than for a corresponding volume of pure dielectric, with higher potential gradients and dielectric losses per unit volume of dielectric than would occur in the absence of the conducting particles. SCHNEIDER, in reply,[37] has questioned the validity of the simplified configurations considered by HOWE and NISSEN and repeats the claim to experimental support for his statements without, however, giving any details of the experimental evidence. In the absence of such experimental data the contribution of the dielectric matrix to the total losses cannot be assessed with certainty, though it is certain that some dielectric loss will occur as a normal consequence of the presence of the dielectric matrix in the electric field of a coil wound on a core of the composite material. Apart from any particular mode of construction of the material, such losses could be minimised by a suitable disposition of the winding.

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In any experimental analysis of losses it would probably be difficult to distinguish between dielectric and hysteretic loss, since they would be likely to vary in a somewhat similar manner with frequency and amplitude.

9. TOTAL LOSSES IN IRON-POWDER CORE-MATERIALS

Although, as pointed out in section 7, the formation of eddy-currents results in a complex interaction of all the factors involved in the absorption of power in ferro-magnetic cores, CAUER has shown[32] that under practical conditions only a small error is caused by assuming that the eddy-currents do not disturb the distribution of magnetic flux, an assumption which makes it possible to ignore the interaction of eddy-current and hysteresis-loss and to treat them as independent of each other. On this basis a convenient formulation of the total loss in telephone cable loading-coils at audio frequencies, is given by MARRIS, POLGREEN and WILLIAMS[31] as follows :—

R = total effective resistance at frequency f cycles per second of a coil of inductance L henries, wound on a ferromagnetic-powder core.

K = constant depending on winding space and wire gauge.

μ = permeability.

Q_e = eddy-current loss-coefficient at a specified frequency f_1 (2,000 cycles per second unless otherwise stated).

Q_n = coefficient of hystereses-loss at zero flux density and zero frequency (" nachwirkung " or " retardation effect ").

Q_h = coefficient of hystereses-loss (slope of the hystereses-flux density curve).

B = flux density.

$$\frac{R}{fL} = \frac{K}{\mu f} + Q_n\mu + Q_h\mu + Q_hB\mu + \frac{Q_e}{f_1} f\,\mu.$$

This shows R as a linear function of B and a quadratic function of f, and permits of an experimental analysis of the losses by plotting R/fL as a function of f with B as parameter. Ignoring the small term $K/\mu f$, this will give a series of straight lines. A typical example is given in the paper referred to,[31] for frequencies up to 2,000 cycles per second.

There are no published data showing whether this formulation is valid at radio frequencies for cores designed for radio-frequency operation. J. H. MORECROFT[38] gives measured values of resistance up to 100 kilocycles per second for a coil of 275 turns wound on a toroidal iron-powder core, 10 cms. external diameter, rectangular section (2·5 cm. wide by 2 cm. deep), inductance about 8 *m*H.

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The values are as follows :—

f (kc/s.)	10	20	30	40	50	60	70	80	90	100
R (ohms)	5	8	12·5	19	26	38	51	67·5	89	125

From these figures it can be shown that R is represented very approximately by :—

$$R = 2{\cdot}3 + 2{\cdot}6 \times 10^{-1}F + 6{\cdot}5 \times 10^{-4}F^2 + 7{\cdot}9 \times 10^{-5}F^3$$

F being the frequency in kilocycles per second. For this case, therefore, a quadratic formulation is quite inadequate, and the term proportional to the cube of the frequency predominates at the higher frequencies. (At 100 kilocycles the successive terms are 2.3, 26, 6.5, 79). An approach to resonance will not account for this behaviour, the tuning capacity for 100 kilocycles per second being over 300$\mu\mu$F. A possible explanation is an increase with frequency of the hysteresis-loss per cycle.

This case, however, is not necessarily representative of the behaviour of radio-frequency core-materials, for the core used had a permeability of about 50, which is higher than would be suitable for radio-frequency use.

The coil-inductance is shown as being absolutely constant up to 40 kilocycles per second, beyond which point there is actually a small increase, rising to about 13 per cent. at 100 kilocycles per second. The most probable explanation for this is the ordinary effect of self capacity on the apparent inductance of a coil. In the absence of data on this feature of the coil, the nature and degree of the variation of permeability with frequency cannot be deducted, but it is improbable that there is any large variation of permeability over the frequency range of the observations.

9.1. The Losses Expressed as a Power-Factor of the Core-Material

It is clearly desirable that the electrical behaviour of ferro-magnetic core-material shall be expressed in terms of specific characteristics of the material itself. In this respect the present subject may be compared with the investigation of the behaviour of dielectrics, the most important electrical characteristics of which are specified as the permittivity and power-factor of the dielectric medium. The analogous terms in the present case would be the permeability and power-factor of the permeable (or ferro-magnetic) medium. The term does not appear to have been much used in the present connection, but Wolman[4] and Deutschmann[5] make use of the equivalent conception of complex permeability.

The idea has already been introduced (in section 7) but it will be well to consider it rather more carefully in its practical significance.

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In the case of dielectric media, measurements can be designed so that the determination of the capacitance and resistance of a test-condenser can be expressed precisely in terms of the permittivity and power factor of the test-specimen of the dielectric medium. In a similar manner, the permeability and power-factor of a given ferro-magnetic core-material is theoretically determinable from measurements of the inductance and resistance of a coil wound upon a core of the material. The resistance of such a coil at a frequency f $(= \omega/2\pi)$ will consist of two main terms R_0 and R, representing the copper and magnetic-core losses respectively. Similarly the inductance can be considered to consist of two main parts L_0 and L due to leakage flux not passing through the core and the induction in the core itself. For a given sinoidal current i the potential will be similarly divisible into two parts v_0 and v, where

$$v_0 = (R_0 + j\omega L_0)\, i$$

and

$$v = (R + j\omega L)\, i$$

(It must be emphasised at this point that this formulation implies that R and L are at least sufficiently independent of i (*i.e.*, of flux density) to make it permissible to neglect the possible wave-form distortion due to the magnetic core. This is, however, a necessary condition for the satisfactory application of such material to radio-reception).

It is desired to express the second component of the voltage, *i.e.*, the component which is directly associated with the core, in terms of the characteristic properties of the core. This can be done by assigning a complex permeability

$$\mu_e = \mu - jr$$

to the core material. The inductance, assumed proportional to the permeability, then becomes itself a complex quantity, *i.e.*,

$$L_e = k\mu_e = k\,(\mu - jr)$$

and, putting

$$v = j\omega L_e i$$

we have

$$v = j\omega k\,(\mu - jr)\, i$$

$$= (k\omega r + j\omega k\mu)\, i$$

and the actual quantities R and L are given in terms of the assumed characteristics of the permeable medium by

$$R = k\omega r$$

$$L = k\mu.$$

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The significant properties of the permeable medium are thus represented by an actual permeability μ and a power factor $\cos \phi$ where

$$\tan \phi = \frac{r}{\mu}$$

Moreover, since

$$\frac{r}{\mu} = \frac{k\omega r}{k\omega\mu} = \frac{R}{\omega L}$$

this power-factor can be determined by a measurement of R and L.

The physical significance of the r term can be demonstrated by consideration of a typical case, for example, a toroidal coil of N turns wound on a core of which the magnetic path is of length l and cross section A. For this case

$$k = \frac{4\pi N^2 A}{10^9 l}.$$

The total power-loss in the core is given by

$$RI^2 = \omega r k I^2 = \text{or } \omega r \frac{4\pi N^2 A}{10^9 l} I^2 \text{ watts.}$$

Since, for the case considered

$$H = \frac{4\pi NI}{10l}$$

$$RI^2 = \frac{r\omega}{4\pi} \frac{16\pi^2 N^2 I^2}{100^2} lA\ 10^{-7} \text{ watts}$$

$$= \frac{rf}{2} H^2 lA\ 10^{-7} \text{ watts.}$$

If W = energy loss in core in ergs per c.c. per cycle.

Power loss in core = $WlAf$ ergs per second.

$= WlAf\ 10^{-7}$ watts,

i.e.,

$$RI^2 = WlAf\ 10^{-7} = \frac{rH^2}{2} lA\ f10^{-7}$$

$$\therefore W = \frac{rH^2}{2},$$

or

$$r = \frac{2W}{H^2}.$$

and

$$\omega r = \frac{8\pi}{H_m^2} Wf.$$

where H_m is the amplitude or maximum value corresponding to the R.M.S. value H.

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From the above relations μ, r can be determined by measurement of L, R and calculation of k.

It will be noted that unless W (*i.e.*, the specific core-loss in ergs per cycle) is proportional to the square of the magnetising force, r (and consequently R) will not be independent of current amplitude. This is obviously a matter of importance in relation to radio-frequency application.

10. THE BEHAVIOUR AND CHARACTERISTICS OF SOME EXISTING TYPES OF IRON-POWDER CORE-MATERIAL

An article by SCHNEIDER[23] gives some account of the construction of an iron-powder core-material recently developed in Germany. A magnetic method of particle-alignment and a lamination of the core by interleaving with thin paper are described as distinctive features of this process. It is suggested that certain dielectric losses of appreciable amount are minimised by these features, but this claim has been questioned by HOWE[34] and NISSEN.[35]

A paper by FRÜHAUF[24] gives the results of measurements of the power-factors of coils wound toroidally on cores of this material, in comparison with air-cored solenoids of the same diameter but about three times the axial dimension.

A typical set of results is as follows :—

A = Toroidal coil with iron-powder core. B = Air-cored solenoid.

λ in metres	150	300	400	500
R in ohms ..	A ..	40	6	3	2
	B ..	12	5	3	3
Rejector impedance, ohms	A	80,000	110,000	110,000	100,000
	B	180,000	125,000	80,000	75,000

It will be noted that the magnetic-cored coil is of lower tuned rejector-impedance than the other at the shorter wavelengths, but higher at the longer wavelengths, owing to the fact that this quantity is practically constant for the magnetic-cored coil but falls fairly rapidly with increasing wavelength in the air-cored coil. The practical constancy of the magnetic-cored coil in this respect is consistent with the theory outlined in section 7, which shows that the eddy-current losses are approximately equivalent to a constant shunt resistance.

When the air-cored coil is screened by a metal case of about twice the coil diameter, and the magnetic-cored coil by a close screen, the comparison is more in favour of the latter, since its resistance is not materially affected by screening and that of the air-cored coil is appreciably increased.

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The same toroidal former with the winding spaced away from the core (C) and wound direct on to the core (D) gave comparative results as follows, for tuned rejector-impedance.

	150	300	400	500	λ in metres.
C	80,000	110,000	110,000	100,000	
D	30,000	65,000	70,000	80,000	

The reduced values in the second case (corresponding to higher coil resistances) are attributed to increased self-capacity due to the proximity to the metallic surfaces of the core.

There appear to be no published data for this material showing the variation of effective coil-resistance with amplitude nor of permeability with flux-density. The normal value of the permeability is about 10. The temperature coefficient has not been stated, but the writer has reason to believe that coils wound on this material will show a relatively large negative temperature-coefficient of inductance.

Another iron-powder core material for use at radio frequencies has been developed in America and is described by POLYDOROFF in an article published in the Proceedings of the Institute of Radio Engineers.[26] (A previous brief description will also be found on p. 914 of Vol. 20 of these Proceedings). The material is stated to consist of chemically-produced iron-particles of about 5–10 microns diameter, with an insulating film of a special varnish, the film being about 1 micron thick. The matrix is a phenol resin or " bakelite," the iron-content of the mass being 95 per cent. The permeability is about 12.

POLYDOROFF states that the permeability is constant with respect to frequency from 50 to 2×10^6 cycles per second, and independence of magnetising force from 0·01 to 10 Gauss. His article is mainly devoted to a description of the application of this material to a method of permeability-tuning having as its object the maintenance of a uniform " selectance " over a given frequency-band. He proposes as a factor of merit for a core-material the quantity $\delta L/\delta R$ (the increments of inductance and resistance as compared with an air core) and states that this is of the order 15×10^{-6} (henries per ohm) in the material he describes. No data are given in the above publication relating to temperature coefficient but the writer has been informed by POLYDOROFF that it is about 2 parts in 10,000 per degree Centigrade. There is no exact information as to the variation, if any, of coil-resistance with amplitude.

A number of iron-powder core materials have been developed in this country but no detailed accounts have as yet been published in the scientific or technical journals. Coil-resistance and inductance measurements for one such material have been included in a pamphlet issued by the manufacturers. No details of composition or manufacture are given, but the permeability is stated to be

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about 4. Comparative coil-resistance data are given for two coils of equal inductance (160 μH) over a range 500-1,500 kilocycles per second. One of these is a single-layer 27/32 Litz-winding on a 3-in. glass former, and the other a solid-wire winding on a toroidal iron-powder core 1½-in. diameter by ½-in. deep. From 500–800 kilocycles per second, the Litz coil gives a slightly higher value of $\omega L/R$, but from 800 to 1,500 kilocycles per second the coils are practically equivalent. The $\omega L/R$ values for the iron-powder cored coil range from 120–80 over the frequency-range 500–1,500 kilocycles per second, and those for the Litz-wire coil from 200–80.

The temperature coefficient is not stated, but the writer has reason to believe that a coil wound on the material will show a positive temperature-coefficient of inductance of the order of one in ten thousand. There appears to be no published information as to the variation, if any, of inductance and resistance with flux-density.

It is desired to emphasise that the above selection of published information is included only as an indication of the possibilities of iron-powder core-material for use at radio frequencies. A comparison of the materials from a practical point of view would involve a number of other factors of a technical and economic character which do not come within the scope of this survey and no such comparison is either intended or implied.

11. SUMMARY AND CONCLUSIONS

It can be taken as established experimentally, that the actual initial permeability of iron and ferromagnetic alloys is independent of frequency at least up to the highest practical radio frequencies.[10, 11, 12] It appears, however, that the actual shape of the μ-H variation may depend on frequency, the tendency being to a smaller variation of μ with H at low values of H.

The effective d.c. and low-frequency permeability of magnetic-powder compositions is low compared with that of the magnetic material and depends mainly upon the particle size and spacing rather than on the permeability of the magnetic constituent.[5, 6, 7, 22]

On the basis of experimental evidence it appears probable that the actual specific hysteresis-loss per cycle in magnetic material increases with frequency. The nature and degree of this variation, if any, are at present not known.[17, 19]

There is in alternating current operation an additional loss factor known as " nachwirkung " or " retardation loss," the magnitude and variation of which at radio frequencies are not known.[5]

It is possible that dielectric losses due to the power-factor of the insulating material in iron powder compositions are significant at radio frequencies[23] but their magnitude and variation at radio frequencies are not known.

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It is probable, though not certain, that the greater part of the total radio-frequency losses in composite magnetic material is due to eddy-current formation and a consequent skin-effect in flux-distribution. On theoretical grounds it appears a necessary condition of economic applicability of magnetic material at radio frequencies that the fineness of division of the magnetic constituent shall be such that the flux distribution remains substantially uniform, under which conditions the apparent permeability of the material will not differ greatly from the d.c. permeability, and the eddy-current losses will be proportional to the square of the frequency.[4]

There are at present a number of iron-powder compositions in which, by sufficiently fine division of the magnetic material and provision of adequate particle-insulation, the total losses are reduced to a value which gives an economic permeability of from 4–10 up to frequencies of about 2,000 kilocycles per second. It is not known how far the behaviour of such material at radio frequencies is consistent with theory, nor is there any certain knowledge of the constitution of the losses. There are no published data of the effect of such materials on wave form as a function of frequency and flux density. There is no evidence that the limit of improvement in such materials has yet been reached.

The useful lines of future investigation are indicated by implication in the above summary.

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